

ARTIFICIAL TRANSMISSION LINES IN
TELEVISION CIRCUITS

HAROLD RAYMOND WALKER

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ARTIFICIAL TRANSMISSION LINES
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ARTIFICIAL TRANSMISSION LINES
IN TELEVISION CIRCUITS

By

Harold Raymond Walker
Lieutenant, United States Navy

Submitted in partial fulfillment
of the requirements
for the Certificate of Completion
In Engineering Electronics

United States Naval Postgraduate School
Monterey, California
1953

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PREFACE

Many television circuits and auxiliary devices have appeared on the market recently which incorporate artificial transmission lines as a basic part of their design. The two main functions served by artificial transmission line sections are impedance matching and capacity isolation, but their use for these functions make possible extremely wide bandwidth and comparatively low noise figures when compared with tuned circuits as ordinarily used.

This thesis develops the mathematics necessary for the design of artificial transmission line sections in the resonant single frequency case and by means of transfer functions extends this to the broadband case as normally used. Much of the work on extending the operating frequency over a broadband is original, as are the transfer functions used. This work is not done in any reference known to the author, although it is done by cut and try methods in most laboratories which work with broadband amplifiers.

Because mathematics do not present a clear and immediate picture of the conditions prevailing, graphs are used to permit rapid calculation by cut and try methods on paper.

In order to limit the scope of this thesis, only L, T and π networks are treated, but where an additional improvement results from the use of M derived or constant K sections, this is pointed out. Since the uses of artificial transmission lines given here are essentially lossless, the case of sections with losses will not be considered.

Following the presentation of the theory involved is a survey of circuits employing this theory at the present time, plus several possible circuits developed by the author which offer desirable features to television and radar equipment.

The writer wishes to thank Professors P. E. Cooper, D. A. Stentz, and G. J. Thaler for their assistance and advice in doing the necessary laboratory work, and for their constructive criticism in the preparation of this thesis.

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TABLE OF SYMBOLS USED

C	-	Capacitance
E_1	-	Input Voltage
E_2	-	Output Voltage
$E(s)$	-	Voltage a function of (s) (Laplace Transform)
f_c	-	Cutoff Frequency
f_r	-	Resonant Frequency
G_2	-	Shunt Conductance
$I_{(n)}$	-	Loop Current
$I_{(n)}(s)$	-	Current a function of (s) (Laplace Transform)
L	-	Inductance
Q	-	Quality factor - See Definition Page 2.
R_1	-	Series Resistance
R_2	-	Shunt Resistance
R_c	-	Equivalent Series Resistance in Condenser used in place of 2 in certain formulas.
R_L	-	Load Resistance
R_p	-	Plate Resistance
(s)	-	A variable used in Laplace Transforms ($j\omega$)
t	-	Time
X_c	-	Capacitive Reactance
X_L	-	Inductive Reactance
Y	-	Admittance
Z	-	Impedance
Z_o	-	Nominal Line Impedance

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CHAPTER I

BASIC THEORY OF MATCHING SECTIONS

The resonant circuit shown in figure (1) has a series resistance R_1 and a shunting conductance G_2 which shall be interchangeably referred to as R_2 . If this circuit is resonant at a frequency determined by L , C and R , the relationship given in equation (1) gives the driving point impedance of the anti-resonant circuit. If no conductance G_2 is assumed, and the only loss element is R_1 , then at resonance the equation (2) holds. We thus have a means of converting from series to shunt resistance. This relationship is subject to further transformation so that equation (3) results. Equations (4) and (5) are taken from transmission line theory and given here so that their similarity will be noted.

It should further be noted that equation (5) is the equation for matching two line impedances by means of a quarter wave stub. If figure (1) is redrawn as in figure (2), the same anti-resonant circuit becomes an L section which will match R_1 to R_2 . The range of values over which the L section will match impedances is theoretically infinite, however, due to the practical aspects of the circuit, losses become excessive if the Q of the circuit exceeds 10, in which case it is better to use several L sections, each with a lower Q .

The relationship of equation (3) is subject to further manipulation to get equations (6) and (7). Equation (7) is that of a phase and amplitude distortionless line. For a single frequency match, the

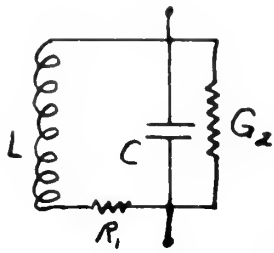


Fig. 1

$$Z_r = \frac{L}{CR} \quad (1)$$

$$R_2 = \frac{L}{CR_1} \quad (2)$$

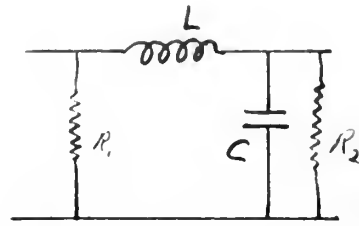


Fig. 2

$$R_1 R_2 = \frac{L}{C} \quad (3)$$

$$Z_o = \sqrt{\frac{L}{C}} \quad (4)$$

$$\sqrt{Z_1 Z_2} = Z_o \quad (5)$$

$$\frac{L}{C} = \frac{\omega L}{\omega C} = X_L X_C = Z_o^2 \quad (6)$$

$$R_2 = \frac{L}{CR_1} \quad R_2 C = \frac{L}{R_1} = \frac{C}{G} \quad (7)$$

$$Q = 2\pi \frac{\text{energy stored}}{\text{energy lost}} \text{ (per cycle)} \quad (8)$$

$$\frac{R_2}{X_C} = \frac{X_L}{R_1} = Q \quad (9)$$

$$X_C = \frac{R_2}{Q} \quad X_L = R_1 Q$$

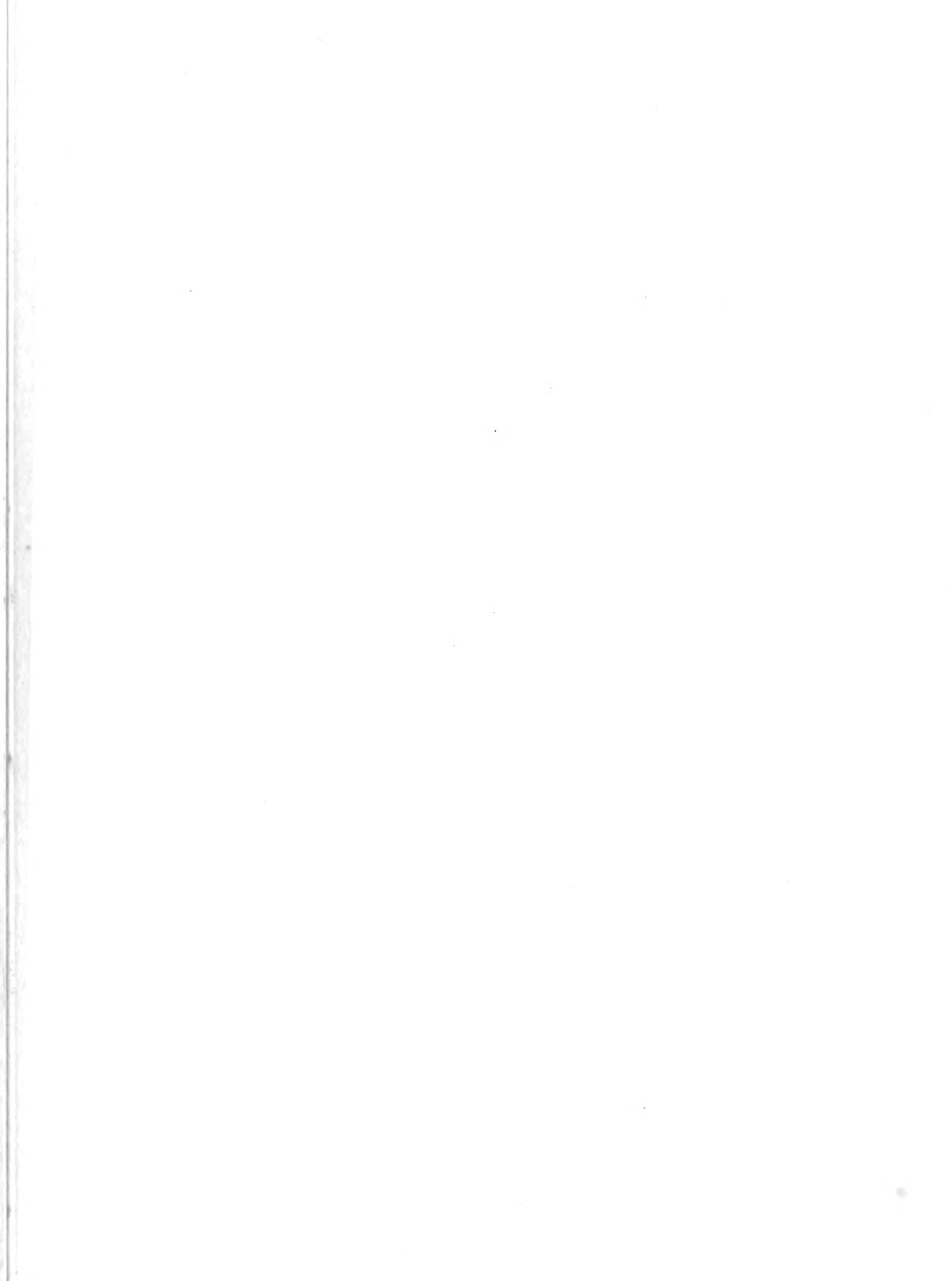
$$\frac{R_2}{X_C} = \frac{X_L}{R_1} = Q \quad (9)$$

$$R_2 X_C = R_2 X_L + R_1 X_C$$

Dividing thru by $X_C R_1$

$$\frac{R_2 X_C}{R_1 X_C} = \frac{R_2 X_L}{X_1 X_C} + \frac{R_1 X_C}{R_1 X_C}$$

$$\frac{R_2}{R_1} = Q^2 + 1 \quad (10)$$



apparent solution is to pick X_L , X_C and Z_0 all equal. Under those conditions resonance is assumed and the input and output should be pure resistive components. For low Q matches, ie., matches where R_1 and R_2 are not widely different, X_L cannot equal X_C as will be apparent from a further investigation of resonance.

The resonant circuit has a Q defined by equation (8). Q may also be defined as the ratio of reactance to resistance as given by equation (9), and as the tangent of the phase angle of current delay or advance. Manipulating equation (9) we obtain equation (10) which is the basis of low Q matching.

Figure (3) gives a universal matching chart for L matching R_1 to R_2 . It will be noted from this chart that the diagonal co-ordinate is Q, and that as Q approaches zero, the capacitive reactance required for a match is greater than the inductive reactance. The reason for this becomes apparent when the basic resonance equation is expanded. Defining resonance as the condition of unity power factor, or pure resistive components, the general resonance formula is given by equation (11). Under conditions of high Q's, R_1 and R_C are small, and the approximation, equation (12) is employed. R_C is the series equivalent of the shunt resistor R_2 which would be assumed to be a resistor in series with C. For most applications other than this formula, it is more convenient to leave it as a conductance G, which represents leakage, tube input conductance, or just a resistive load.

As an example of how Q affects L and C in the L matching circuit, let $Q = 1$. When $Q = 1$, R_2 is two times R_1 by equation (10), and

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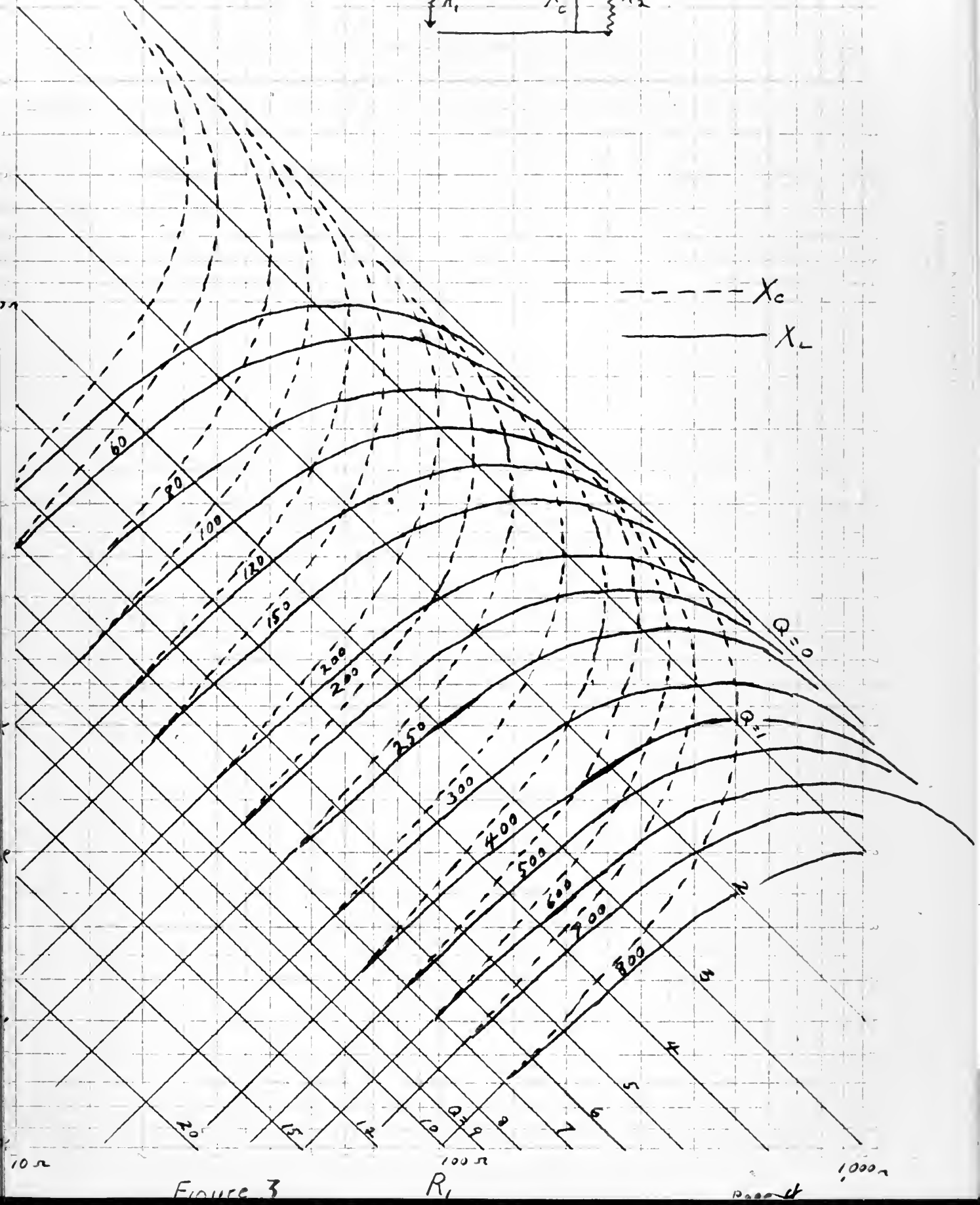
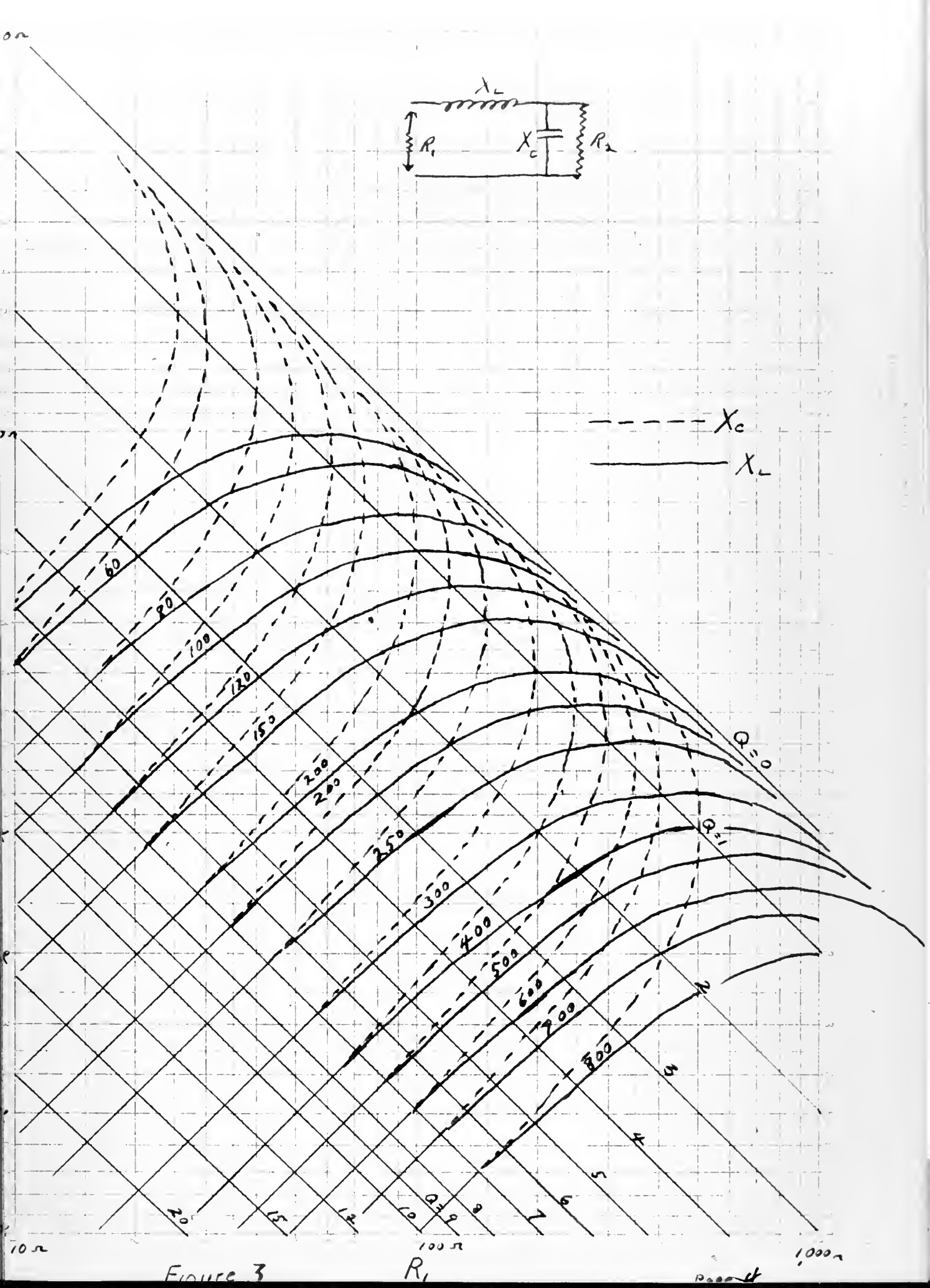
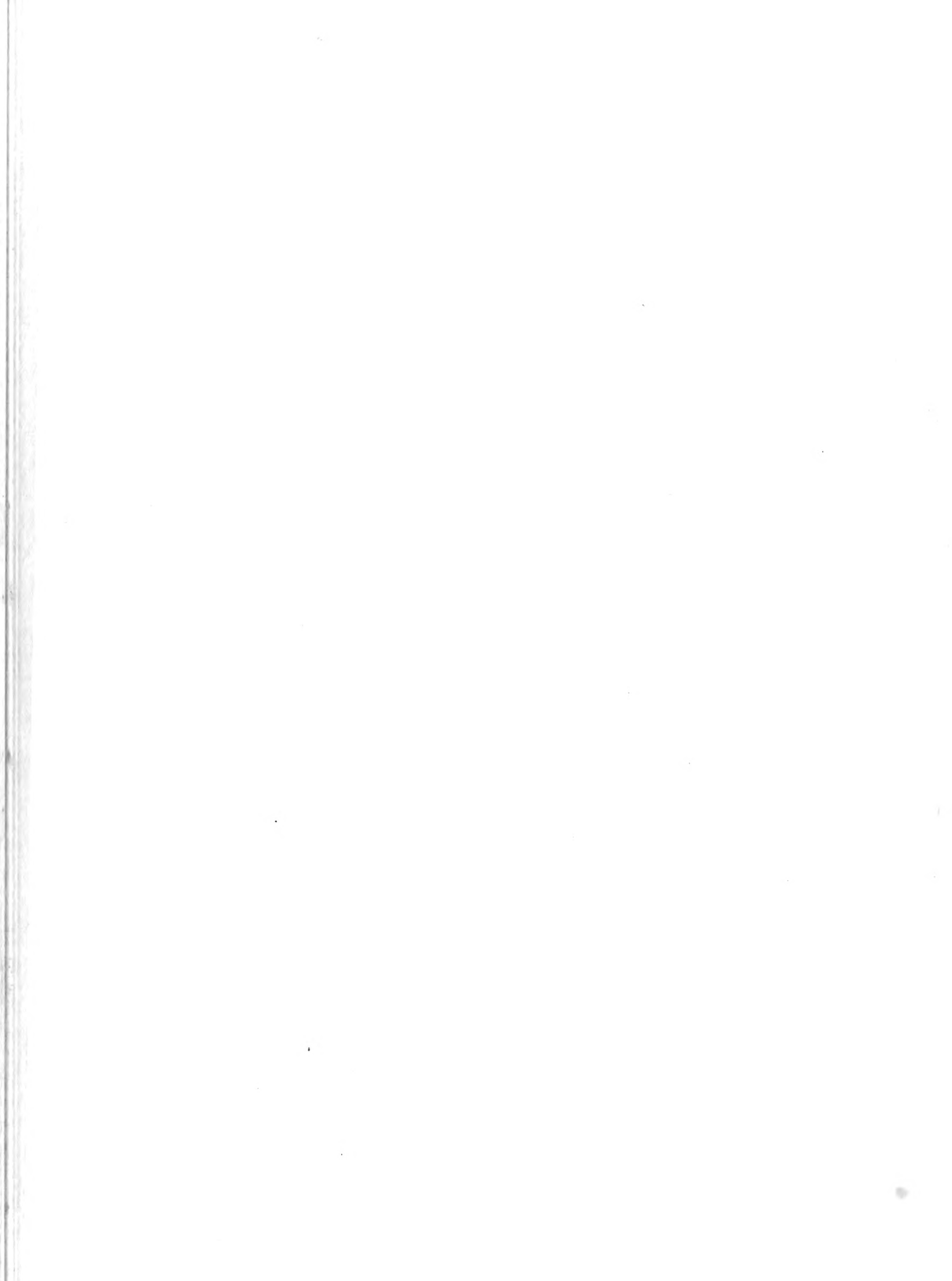


Figure 3

R_1

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similarly R_C is two times R_1 . Taking the large radical in equation (11), substituting equation (13), we obtain equation (14) which indicates the actual resonant frequency is only .707 that indicated by equation (12). Also, since equation (10) states $R_2 = 2R_1$, then X_C must equal $2X_L$ if equation (9) is to be satisfied. It can be seen from figure (3) that $X_C = 2X_L$ when $Q = 1$.

When a chart, such as figure (3), is not available, a convenient rule of thumb using equation (15) may be set up. To use this equation determine Q by equation (10) from the known impedances to be matched. Pick X_C from equation (9) and then pick an X_L which will equal X_C at the frequency given by equation (15). Thus for $Q = 1$, X_L would equal X_C at a frequency 1.4 times the frequency of perfect match. Failure to make the low Q reactance correction results in a loss of about 2 db in most cases, but the match then becomes reactive and may result in distortions in other parts of the circuit.

The basic L section is only a step in the complete matching process. Often it may be enough, but generally several sections are required either front to front as pi sections, end to end as a ladder network, or back to back as a T section. If R_1 is made equal to R_2 , $Q = 0$ and R_2 can be replaced by another similar L section. When this is done, several times, an artificial transmission line terminated in Z_0 results. Figure (4) illustrates these various combinations.

It should be noted that the artificial line is symmetrical, and thus has the same transmission characteristics, both ways. Further, any mismatch in Z_0 will cause impedances to reflect from point to point and the

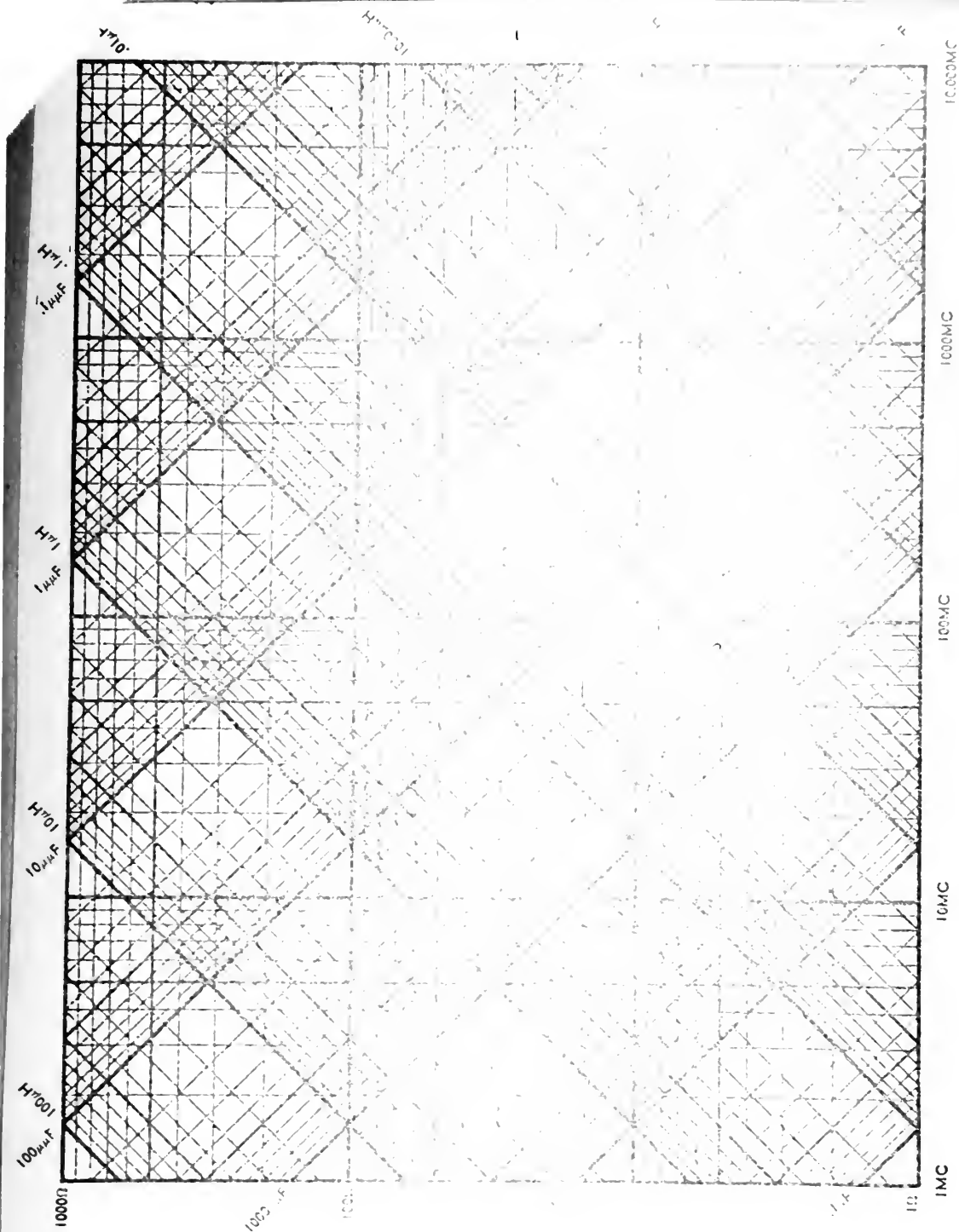


Figure 6

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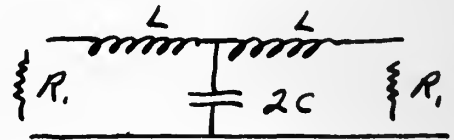
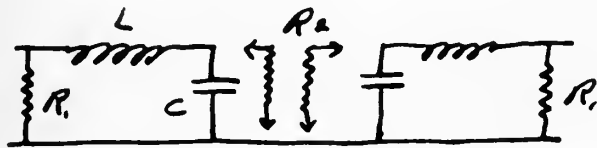
$$(11) \quad T_n = \frac{1}{2\pi\sqrt{LC}} \sqrt{\frac{R_L^2 - \frac{1}{C}}{R_c^2 - \frac{1}{C}}} = \frac{1}{2\pi\sqrt{LC}} \sqrt{\frac{R_L^2 - Z_0^2}{R_c^2 - Z_0^2}}$$

$$(12) \quad f \approx \frac{1}{2\pi\sqrt{LC}}$$

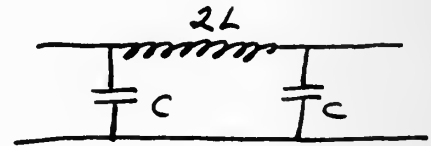
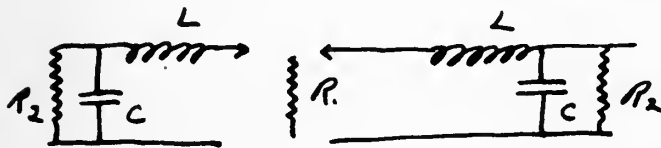
$$(13) \quad Z_0 = \sqrt{\frac{L}{C}} = \sqrt{2} R_1$$

$$(14) \quad f_n = \frac{1}{2\pi\sqrt{LC}} \sqrt{\frac{R_1^2 - (\sqrt{2}R_1)^2}{2R_1^2 - (\sqrt{2}R_1)^2}} = \frac{1}{2\pi\sqrt{LC}} \sqrt{\frac{R_1^2 - 2R_1^2}{4R_1^2 - 2R_1^2}} = \frac{.707}{2\pi\sqrt{LC}}$$

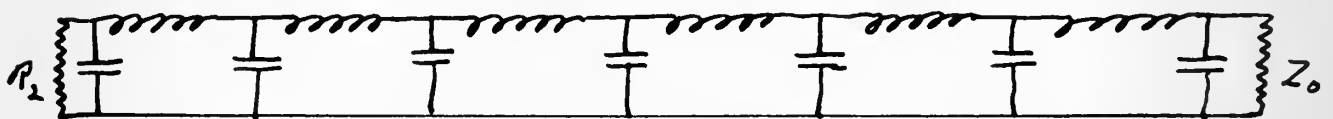
$$(15) \quad f_z = \sqrt{1 + \cot^2 \phi} f_n = \sqrt{1 + \frac{1}{\tan^2 \phi}} f_n = \sqrt{1 + \left(\frac{1}{2Q}\right)^2} f_n$$



equivalent T



equivalent π



artificial line or ladder equivalent

Figure 4

artificial line will reflect and form standing waves in exactly the way they are formed on an actual transmission line. In some of the circuit examples given later the line must be kept flat, in others the line must have standing waves to function.

Figure (4) shows a resistive load at both ends of the section. It should be kept in mind that one of these is the load, the other the driving point impedance. Only one represents a load, the choice of which is up to the designer.

Appendix A-I gives examples of the use of the chart (Figure (3)). Figure (5) is a part of this appendix. Figure (6) is given here as an auxiliary chart to figure (3).

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CHAPTER II

GRAPHICAL SOLUTIONS

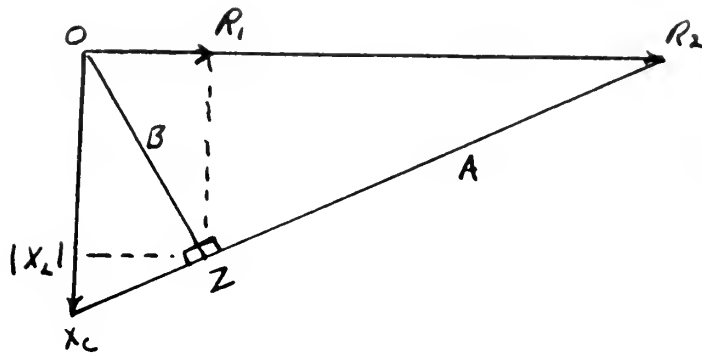
Several graphical solutions to the problem are possible. These are found in a variety of references, chiefly the articles by Paine, Bruene and Storch. The use of the Smith and Bode charts is somewhat original.

The basis of all geometric solutions is to convert parallel to series equivalent circuits by means of similar triangles, and then add series elements vectorially. In figure (7), a parallel RC circuit is assumed, and the reactances plotted out. The reactances are connected by the construction impedance line A and a perpendicular is dropped to it from the origin. The calculations with figure (7) show the line B to be the magnitude and phase of the equivalent series combination.

If the equivalent values are to be used in calculations and not necessarily as a part of the plot necessary for further graphical solution, the solution can be made on a Smith circle chart with normalized values. Use of this chart for this purpose will be taken up later.

Another graphical solution which may be needed, particularly in obtaining transfer functions, is multiplication and division of complex numbers. Appendix A-II shows a graphical method of accomplishing this which may simplify calculations where a large number of values must be tried.

Referring to figure (9(a)), the procedure for setting up the graphical solution is as follows: Lay out on the horizontal axis the value of R_2 , and on the negative vertical axis, the value of X_C . If then,



$$\bar{Z} = \frac{\bar{R}_c \bar{X}_c}{\sqrt{\bar{R}_c^2 + \bar{X}_c^2}}$$

$$\frac{OZ}{OR_2} = \frac{OX_c}{X_c R_2}$$

$$\text{but } X_c R_2 = \sqrt{(OR_2)^2 + (OX_c)^2}$$

hence

$$\frac{\sqrt{(OR_2)^2 + (OX_c)^2}}{OR_2} = \frac{OX_c}{OZ_T}$$

and

$$\bar{OZ} = \frac{(\bar{OX}_c)(\bar{OR}_2)}{\sqrt{(\bar{OR}_2)^2 + (\bar{OX}_c)^2}} = \frac{\bar{X}_c \bar{R}_2}{\bar{R}_2 + \bar{X}_c}$$

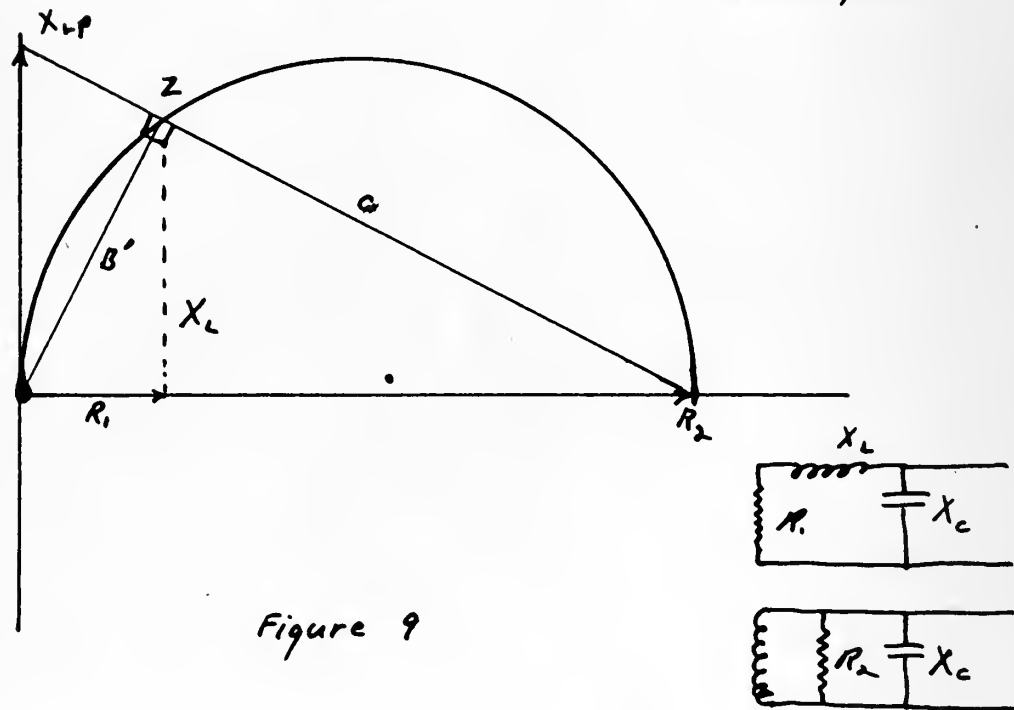
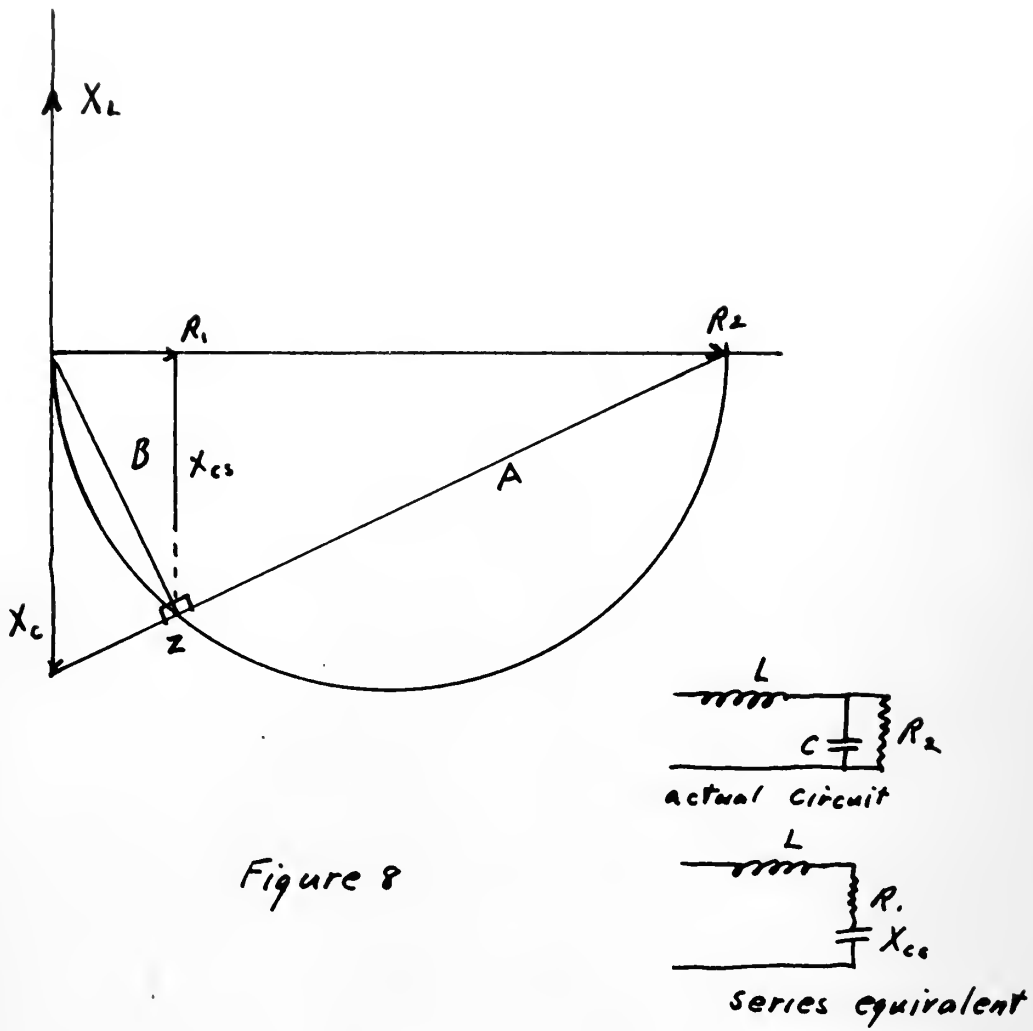
Figure 7.

a circle with diameter equal to R_2 is drawn the intersection of this circle with the line A is the point of the Z vector equivalent to the parallel combination of X_C and R_2 . A perpendicular to this will give the equivalent values of X_{CS} and R_1 as shown. To make a perfect match at one frequency, X_L is chosen equal to X_{CS} . Note that the vector sum of X_L and B then becomes the vector R_1 which has zero power factor.

When X_L and R_1 are known, they are plotted as shown in figure (9(b)). Since the line C must be perpendicular to line B^* , the value of X_{LP} and R_2 can be determined by the intersections of this line with the horizontal and vertical axes, X_C is then made equal to X_{LP} and again the vector sum of B^* and X_C is the pure resistance R_1 .

To graphically solve a pi network, draw two circles of diameter R_2 and R_2^* as shown in figure (10). The value of R_1 is then selected by means of equation (10) from the desired Q. Since R_1 is common, a perpendicular from it to the circles will determine two values of X_L , and the two lines A and A^* will determine the values of input and output capacity. Referring to the figures (9) and (10), the Q of the circuit is represented by the slope of the lines A and C. This being apparent from equation (9). The Q of the pi circuit is modified by the presence of two R_1 values, two X_L values and two X_{CS} values. The overall Q is determined from the sum of these values again plotted and the slope taken as Q, or calculated from equation (9).

Figure (11(a)) shows a graphical plot of the T match. R_1 and R_1^* may be different, but R_2 is a common element. Note that the B and B^* vectors lie on a common circle. From this plot two values of X_C are





as will be shown in the next chapter.

Graphical To Solution
Figure 10

1. The first part of the report is devoted to a general
description of the project and its objectives. It is
followed by a detailed description of the methodology
used in the study. The third part of the report
presents the results of the study, which are
discussed in the fourth part. The final part of the
report is a conclusion, which summarizes the main
findings of the study and provides some suggestions
for further research.

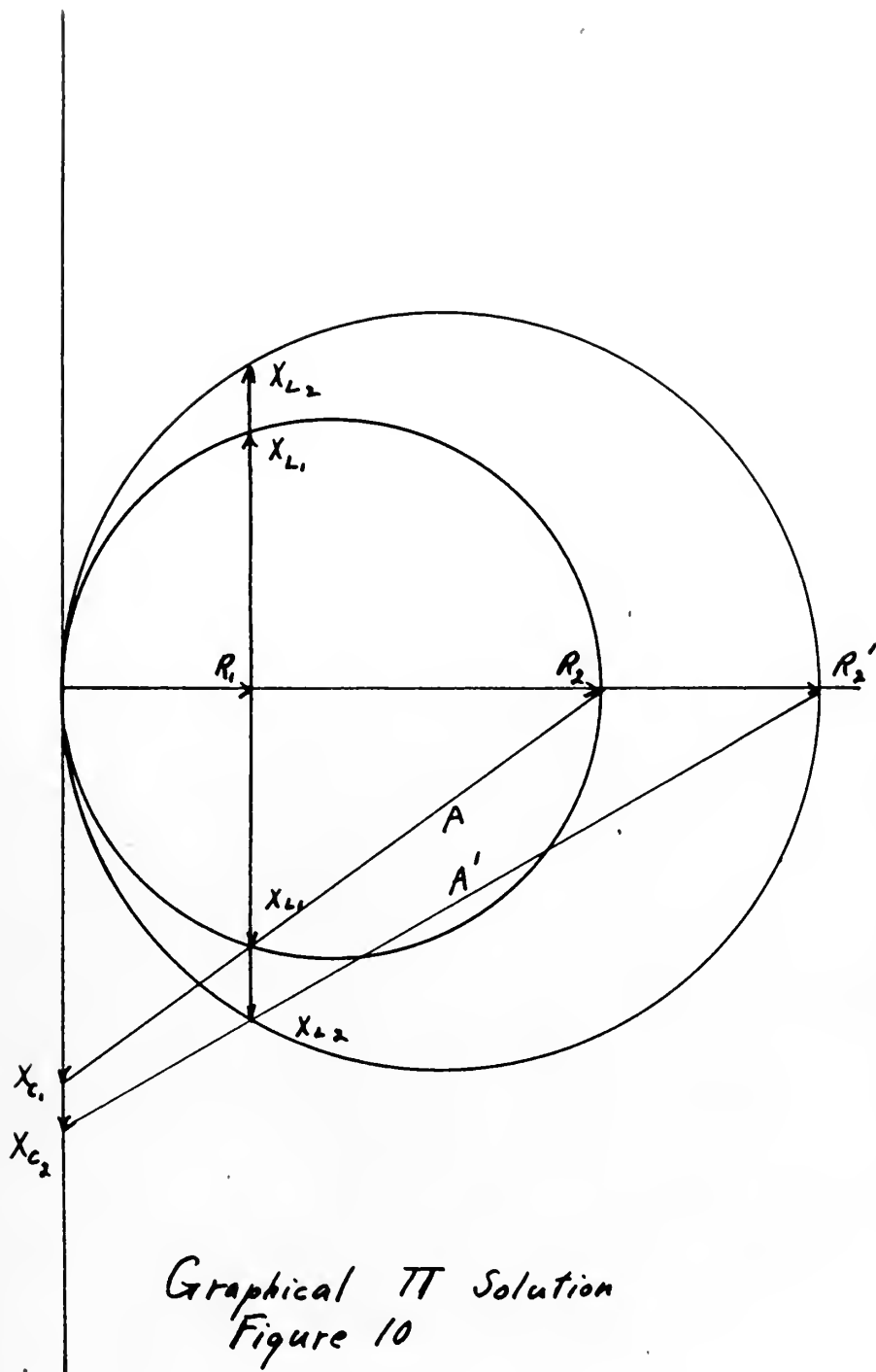
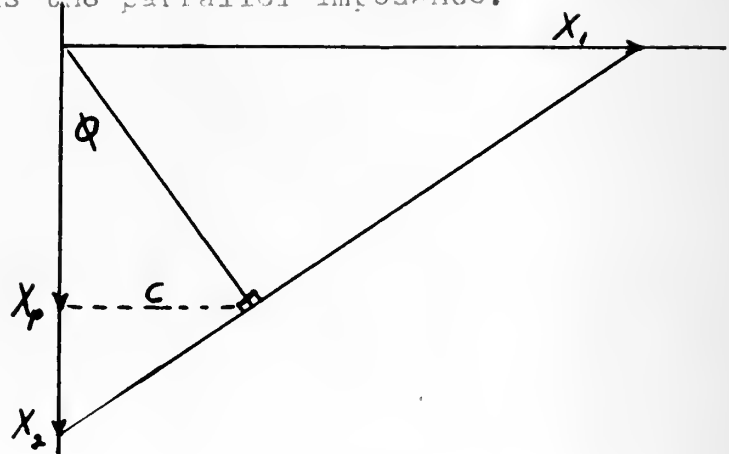
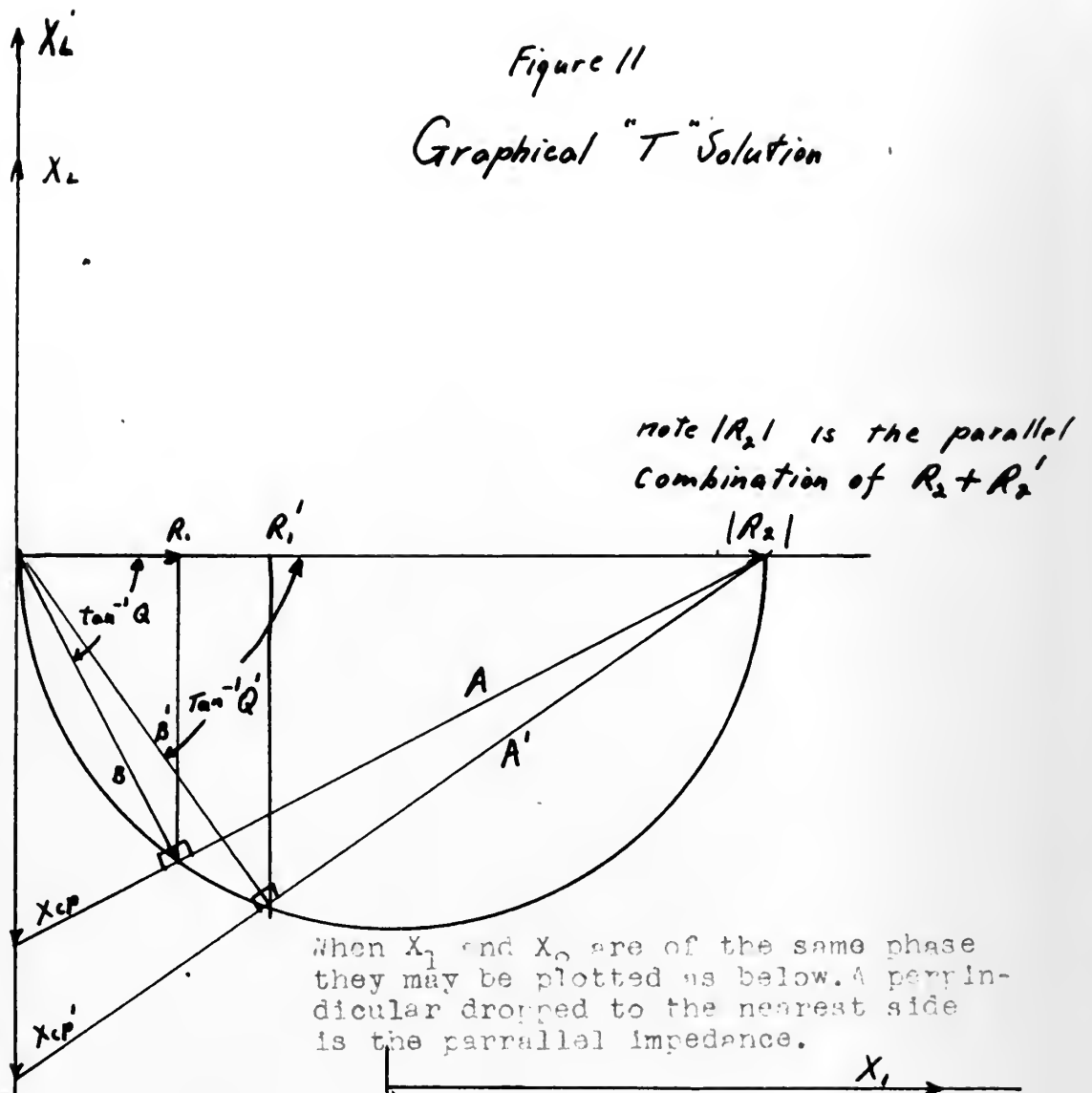
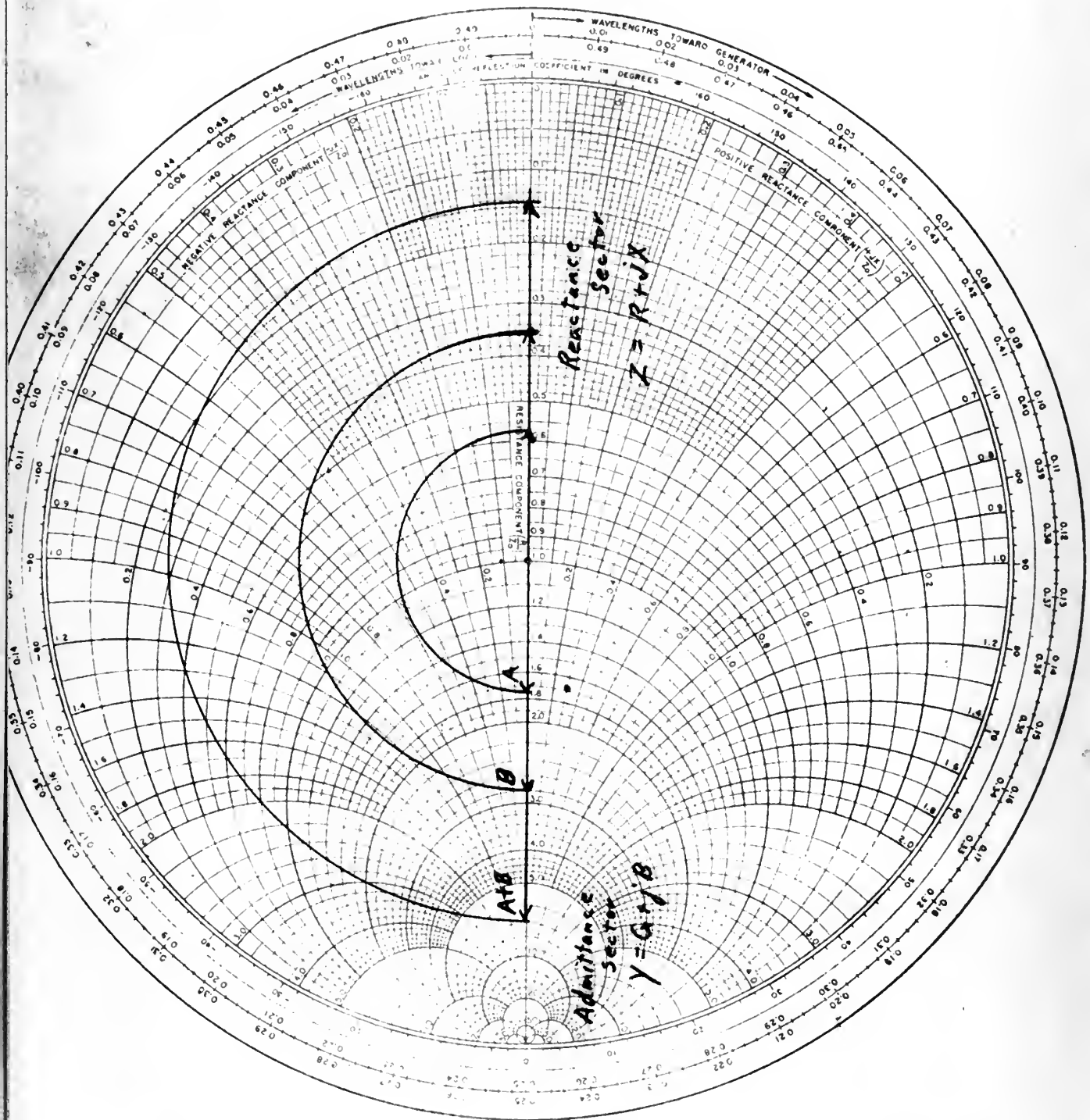


Figure 11
Graphical "T" Solution



Normalize the values of the impedances to 1. Locate the complex impedance on the chart, draw a line thru the center an equal distance out- this locates the complex admittance. Add complex admittances to get the parallel equivalent. No particular sector is for admittance or impedance as shown here-- impedances used in this example are less than 1. Once complex sum is obtained , plot back thru center to get complex impedance.



Smith Chart For Conversion of Series to Parallel Impedances

Figure 12

CHAPTER III

TRANSFER FUNCTIONS OF L, T AND PI NETWORKS

Figure (13) shows the L match. The transfer function is developed as shown to yield equation (16) and a $G(s)$ of the form of equation (16a). The term, (s) , is used to denote Laplace transforms are being used. Most writers on networks employ the term, (p) , which has the same meaning. $G(s)$ is a second order function which will plot in the left hand (s) plane denoting a damped oscillatory circuit. The input impedance has poles at zero and infinity with an intermediate zero. The transfer function has an intermediate pole and a zero at infinity. Whether or not a zero is assumed at zero is a matter of interpretation. The poles and zeros plot as in figure (13).

Figure (14) illustrates the pi match and shows the derivation of its transfer function. The transfer function here is third order denoting a maximum of 270 degrees phase shift and an additional zero in the input impedance.

Figure (15) illustrates the T match and shows the derivation of its transfer function. The transfer function is similar to that of the pi match except that it is multiplied by (s) which indicates a 90 degree phase rotation. Accordingly the poles and zeros plot as shown and appear shifted along the axis.

It will be noted the plots of input Z very closely resemble those of transmission lines of various lengths allowing for a sharp cutoff frequency which limits the number of poles and zeros to those shown.

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... (95) ... (96) ... (97) ...
... (98) ... (99) ... (100) ...

Figures (16), (17), (18), (19), and (20) show the three basic sections plotted for an arbitrary Q . In order to condense the curves, logarithmic scales are used.

Calculating the amplitude and phase variations of the transfer functions is a long laborious process. To condense this time and permit experiments where parameters are varied readily, use was made of the Analog computer. The computer available at the United States Naval Postgraduate School is a 10 amplifier unit built by Boeing Aircraft Corporation. This unit multiplies and integrates, but is unstable when differentiating. For this reason the units are arranged as shown so that no differentiation is required.

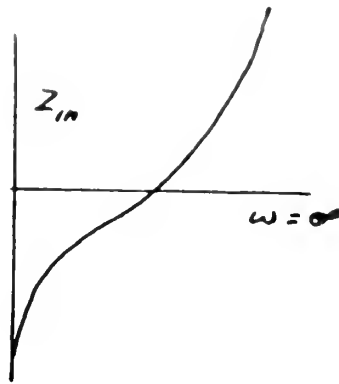
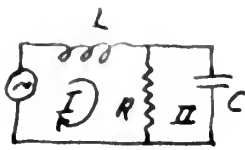
The output is recorded on a two channel brush recorder with a maximum reliable response of about 60 cycles. For this reason the frequency is scaled with a resonance at about 15 cycles. This makes it possible to use readily available R and C values in the computer, and yields a form of universal curve. The values used are $C = 10^{-3}$, $L = .1$ and R according to Q . X_C at 15 cycles is 10 ohms.

The set up for the analog computer used is given in Appendix A-III.

Several conclusions may be drawn from the transfer functions and curves:

1. When sections are cascaded, the resulting transfer function is not $[G(s)]^n$. An artificial transmission line results with standing waves.

2. A special case exists where $Q = 0$, the transfer function is $[G(s)]^n$.



Loop Equations

$$(L_1 s + R_2) I_1(s) - R_2 I_2(s) = E_1(s)$$

$$-R_2 I_1(s) + (R_2 + \frac{1}{Cs}) I_2(s) = 0$$

which yields

$$I_2(s) = \frac{E_1(s) R_2 s}{L_1 R_2 s^2 + (R_2^2 + \frac{L_1}{C}) s + \frac{R_2}{C}}$$

$$E_2(s) = I_2(s) X_C = \frac{I_2(s)}{Cs}$$

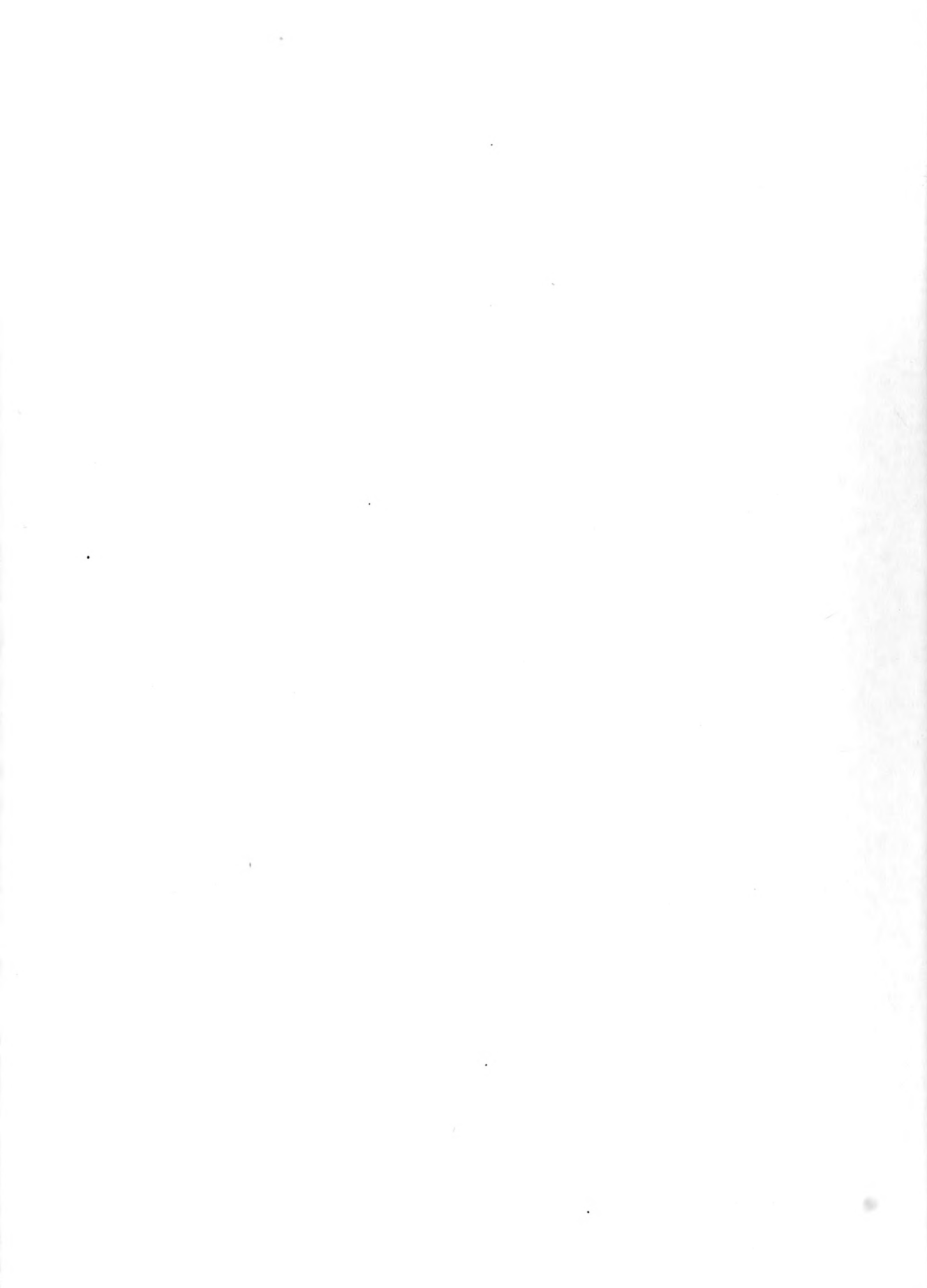
$$\frac{E_2(s)}{E_1(s)} = \frac{R_2}{(\frac{L_1}{C} R_2) s^2 + (\frac{R_2^2}{C} + \frac{L_1}{C^2}) s + \frac{R_2}{C^2}} \quad (16)$$

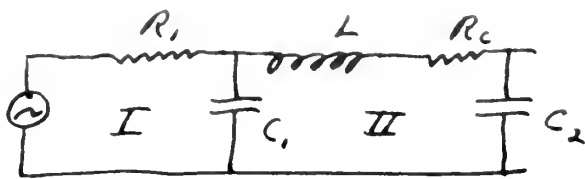
which is of the form

$$G(s) = \frac{a_0}{s^2 + b_1 s + b_0} = \frac{a_0}{(s+\alpha)(s+\beta)} \quad 16(a)$$

Figure 13

Transfer function of L Section



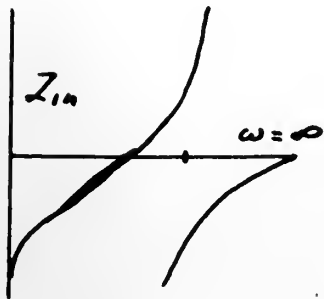


π network driven by source with impedance R_1 . Q is limited by R_c

Loop Equations

$$(R_1 + \frac{1}{C_1 s}) I_1(s) - \frac{1}{C_1 s} I_2(s) = E_1(s)$$

$$-\frac{1}{C_1 s} I_1(s) + (\frac{1}{C_1 s} + \frac{1}{C_2 s} + Ls + R_c) I_2(s) = 0$$



$$I_1(s) = \frac{(\frac{1}{C_1 s} + \frac{1}{C_2 s} + Ls + R_c) I_2(s)}{\frac{1}{C_1 s}}$$

$$= 1 + \frac{C_1}{C_2} + L C_1 s^2 + R_c C_1 s$$

$$(R_1 + \frac{1}{C_1 s}) (1 + \frac{C_1}{C_2} + L C_1 s^2 + R_c C_1 s) I_2(s) - \frac{1}{C_1 s} I_2(s) = E_1(s)$$

$$R_1 + \frac{R_c C_1}{C_2} + R_1 L C_1 s^2 + R_c R_1 C_1 s + \frac{1}{C_2 s} + Ls + R_c = \frac{E_1(s)}{I_2(s)}$$

$$I_2(s) = \frac{E_2(s)}{X_c}$$

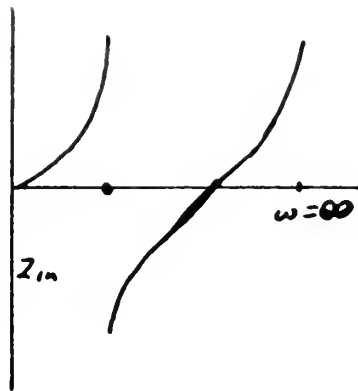
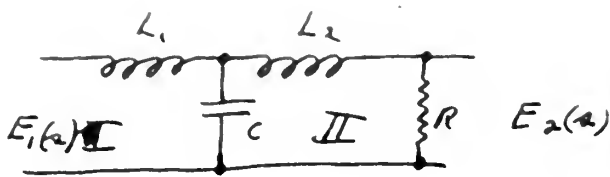
(17)

$$\frac{E_2(s)}{E_1(s)} = \frac{1}{R_1 L C_1 C_2 s^3 + (R_1 R_c C_1 C_2 + L C_1) s^2 + (R_1 C_2 + R_c C_1 + R_2 C_2) s + 1}$$

which is of the form

$$G(s) = \frac{a_0}{s^3 + b_2 s^2 + b_1 s + b_0} \quad (17a)$$

Figure 14- Transfer function of π Section. 20



Loop Equations

$$(L_1 s + \frac{1}{Cs}) I_1(s) - I_2(s) (\frac{1}{Cs}) = E_1(s)$$

$$-(\frac{1}{Cs}) I_1(s) + (L_2 s + R + \frac{1}{Cs}) I_2(s) = 0$$

$$I_1(s) = I_2(s) [Cs(L_2 s + R + \frac{1}{Cs})]$$

$$E_1(s) = I_2(s) [Cs(L_2 s + R + \frac{1}{Cs}) + L_1 s + \frac{1}{Cs}]$$

$$E_2(s) = I_2(s) R$$

$$\begin{aligned} \frac{E_2(s)}{E_1(s)} &= \frac{R}{[Cs(L_2 s + R + \frac{1}{Cs}) + L_1 s + \frac{1}{Cs}]} \\ &= \frac{Rs}{LCs^3 + (RC + L)s^2 + s + \frac{1}{C}} \end{aligned} \quad (18)$$

which is of the form

$$G(s) = \frac{a_1 s}{s^3 + b_2 s^2 + b_1 s + b_0} \quad (18a)$$

Figure 15 - Calculation of $G(s)$ for T Match.

Figure 16

Transfer function of "L" Network
Arbitrary Q

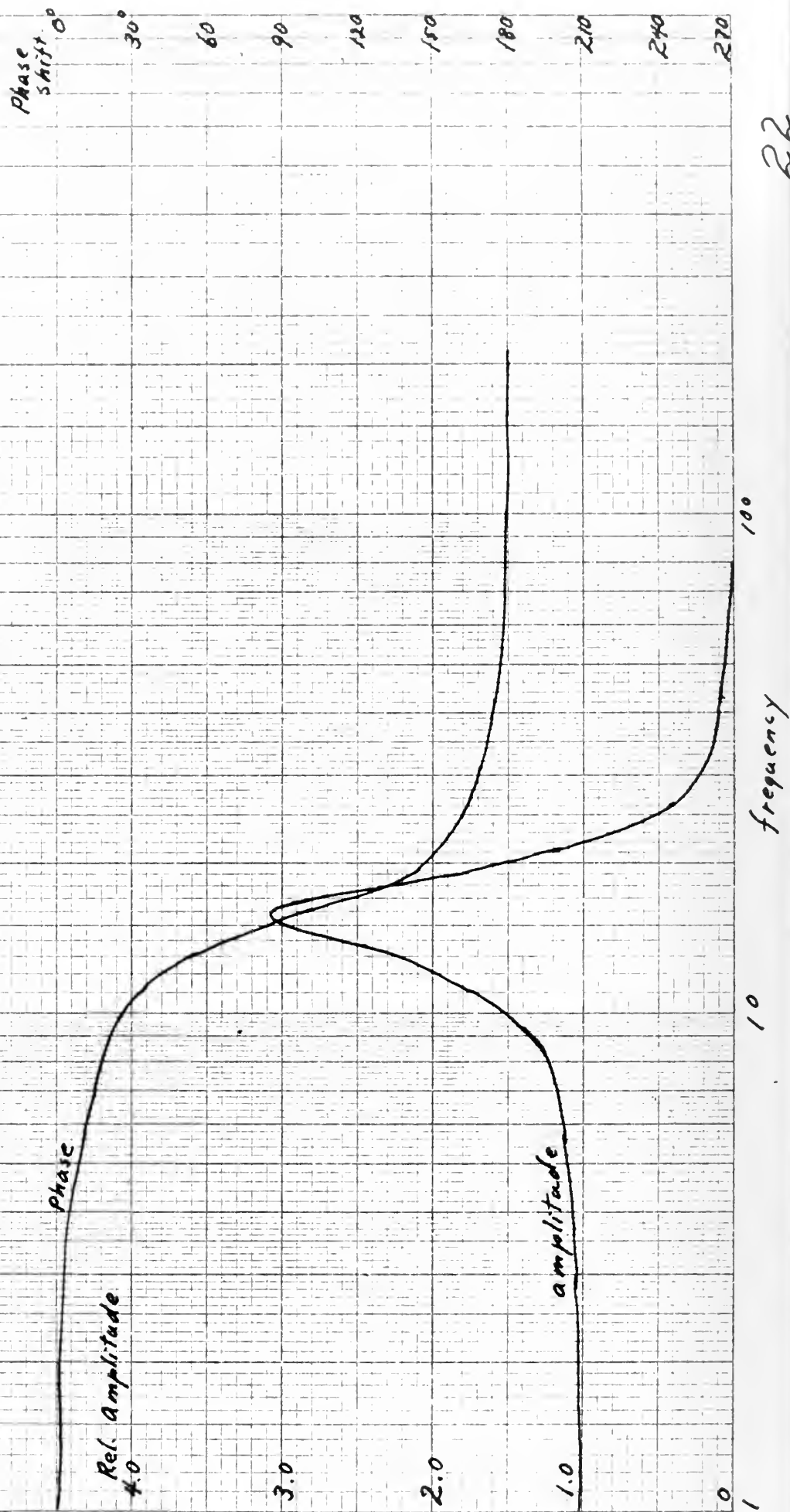
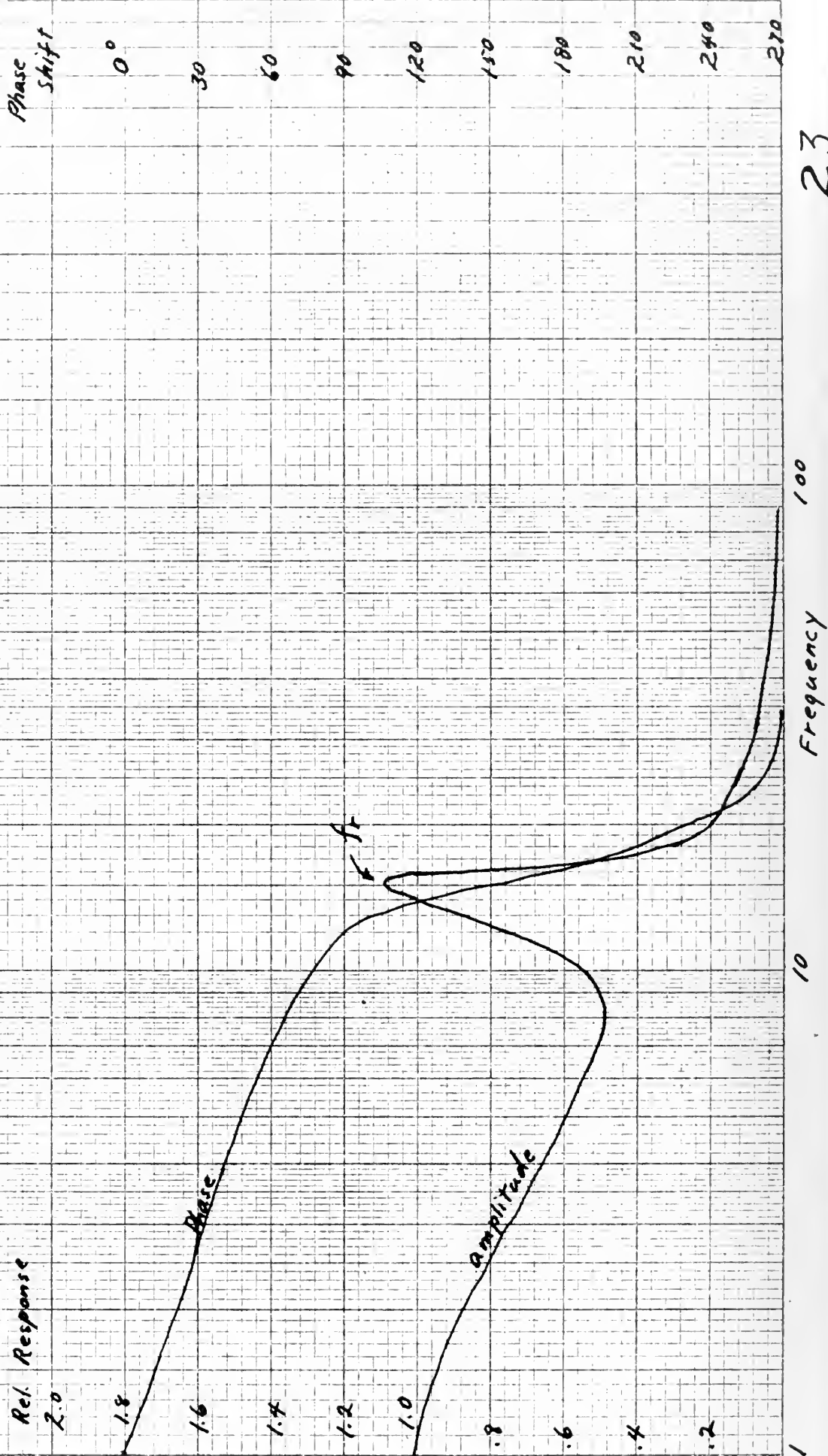


Figure 17

Transfer function of T network
Arbitrary Q



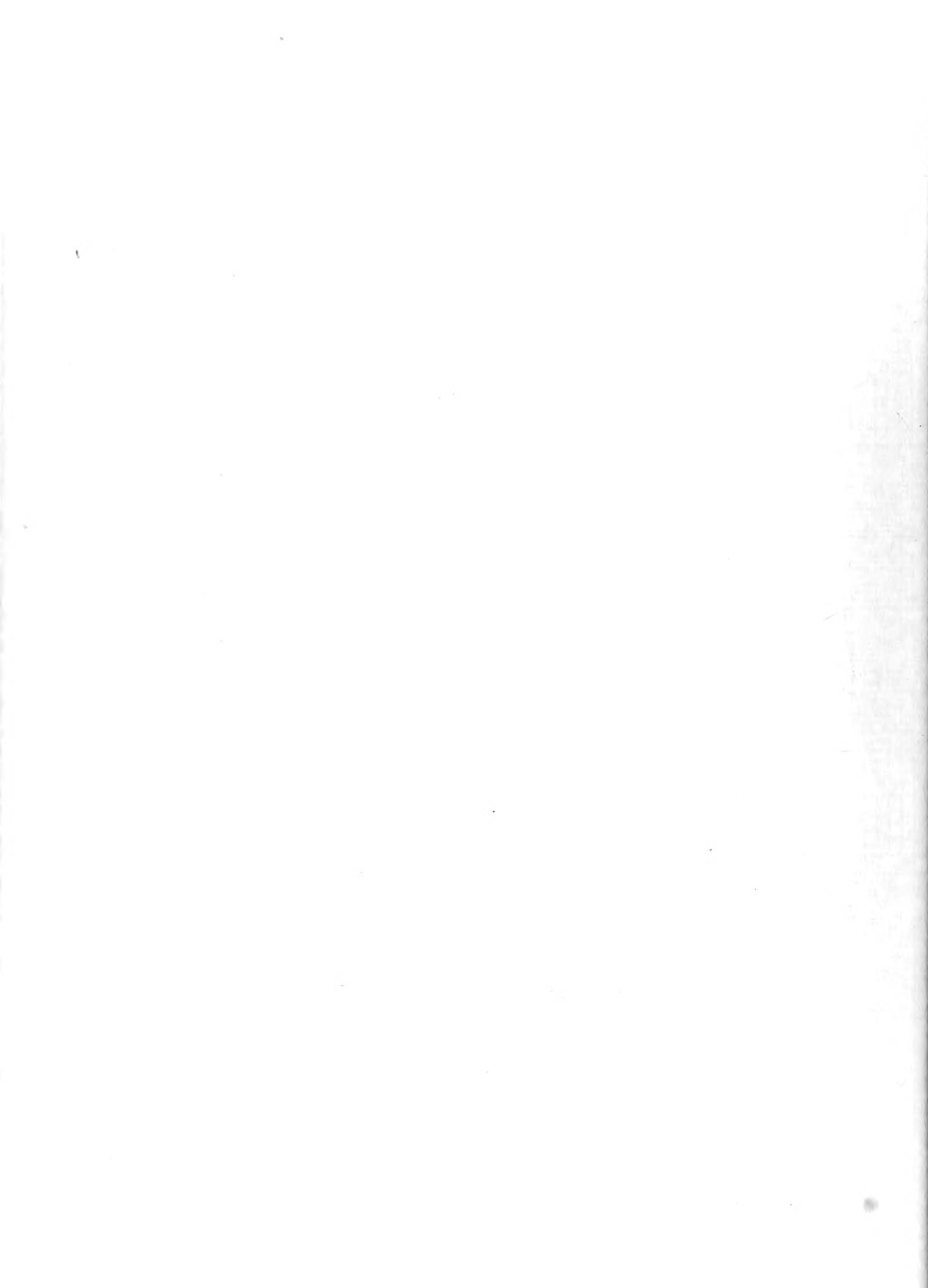
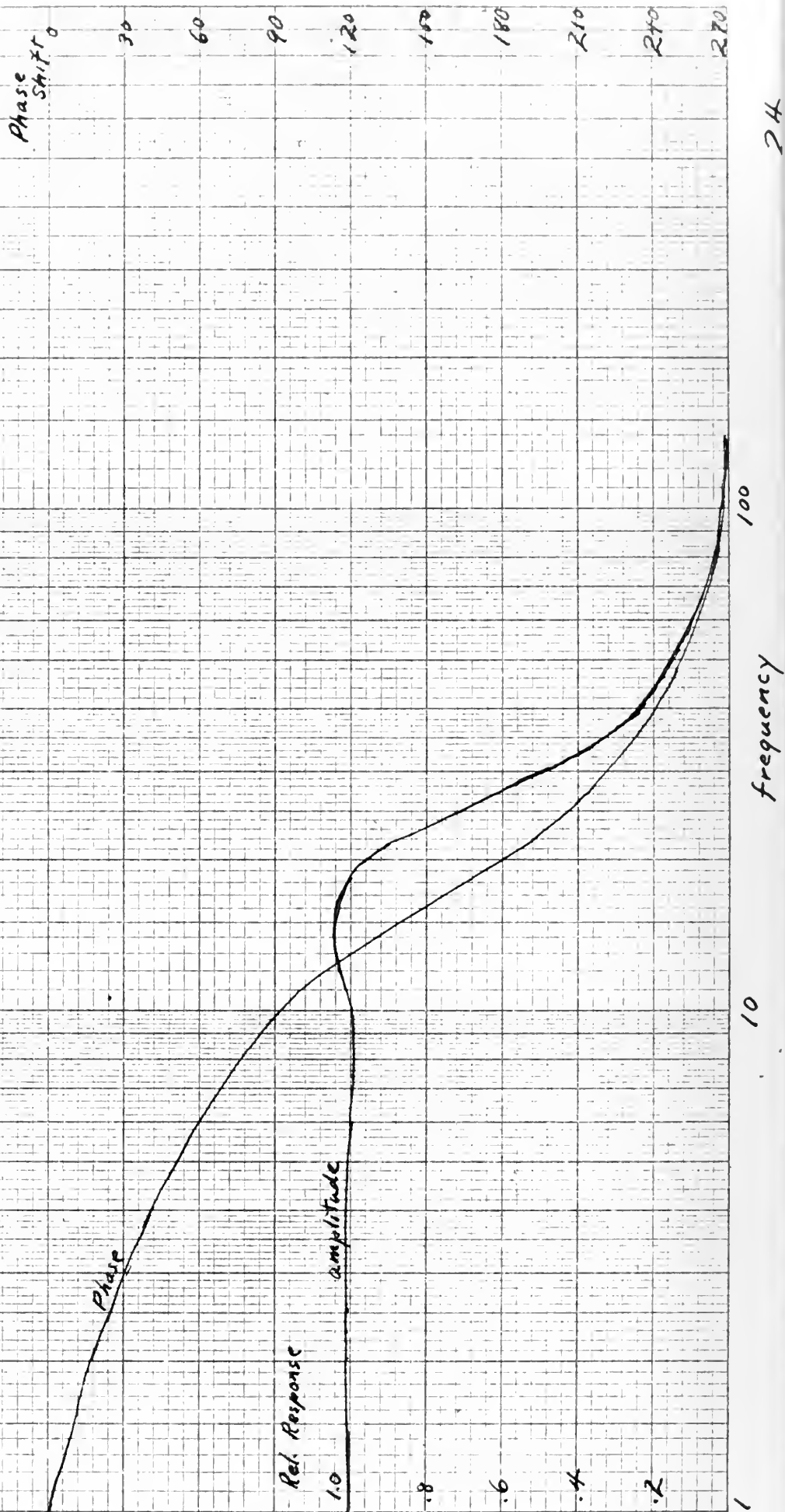


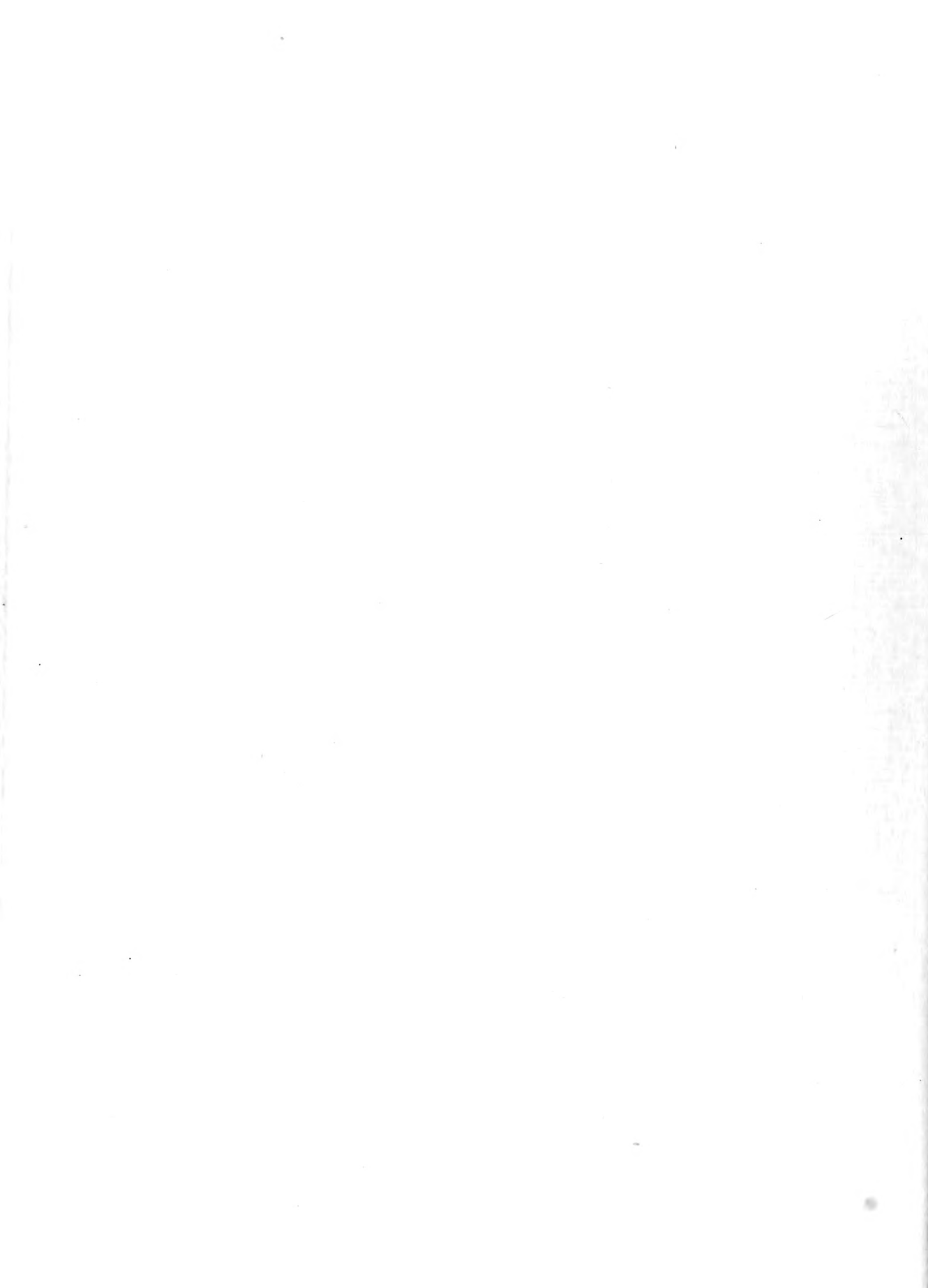
Figure 18

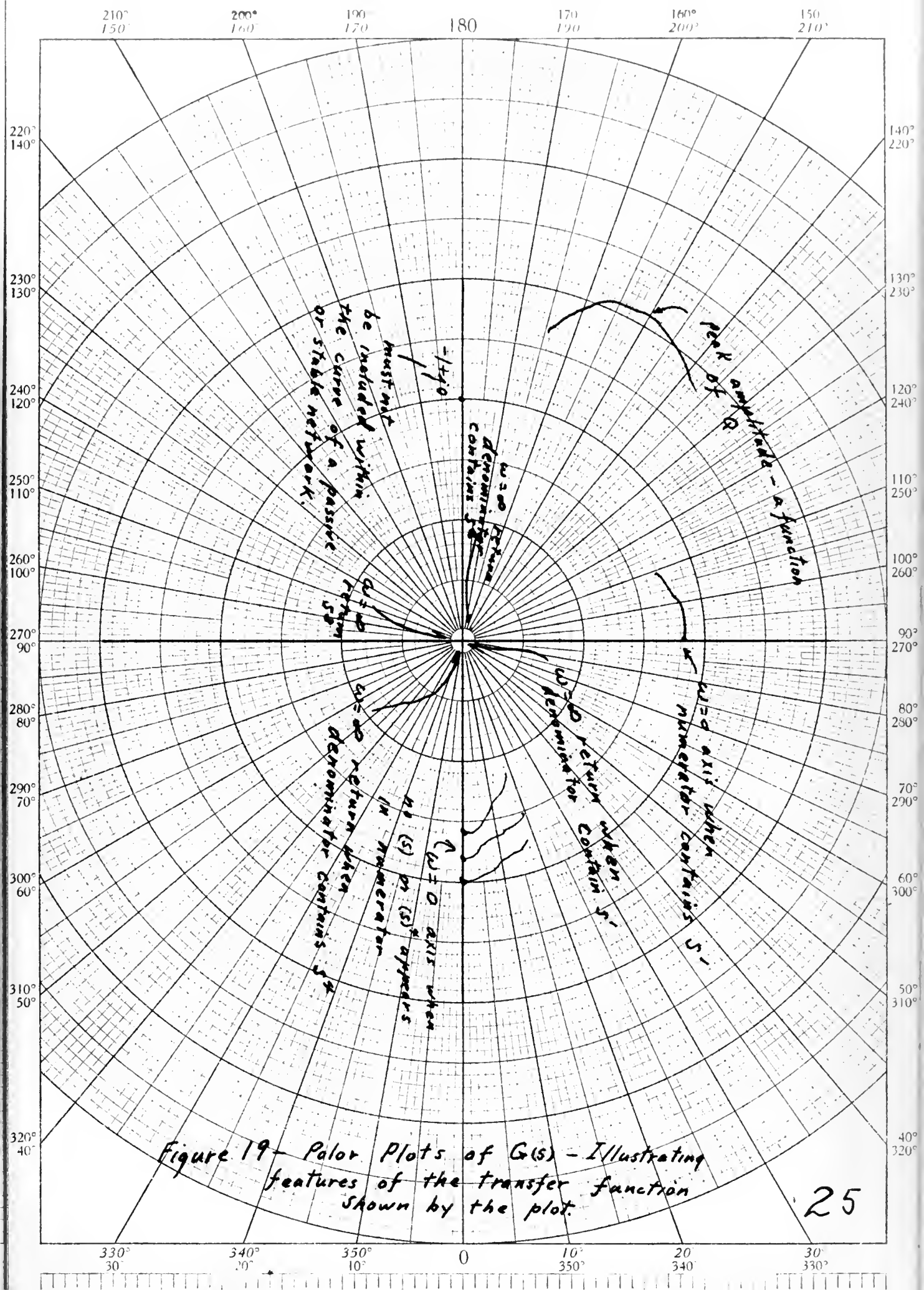
Transfer function of Π network

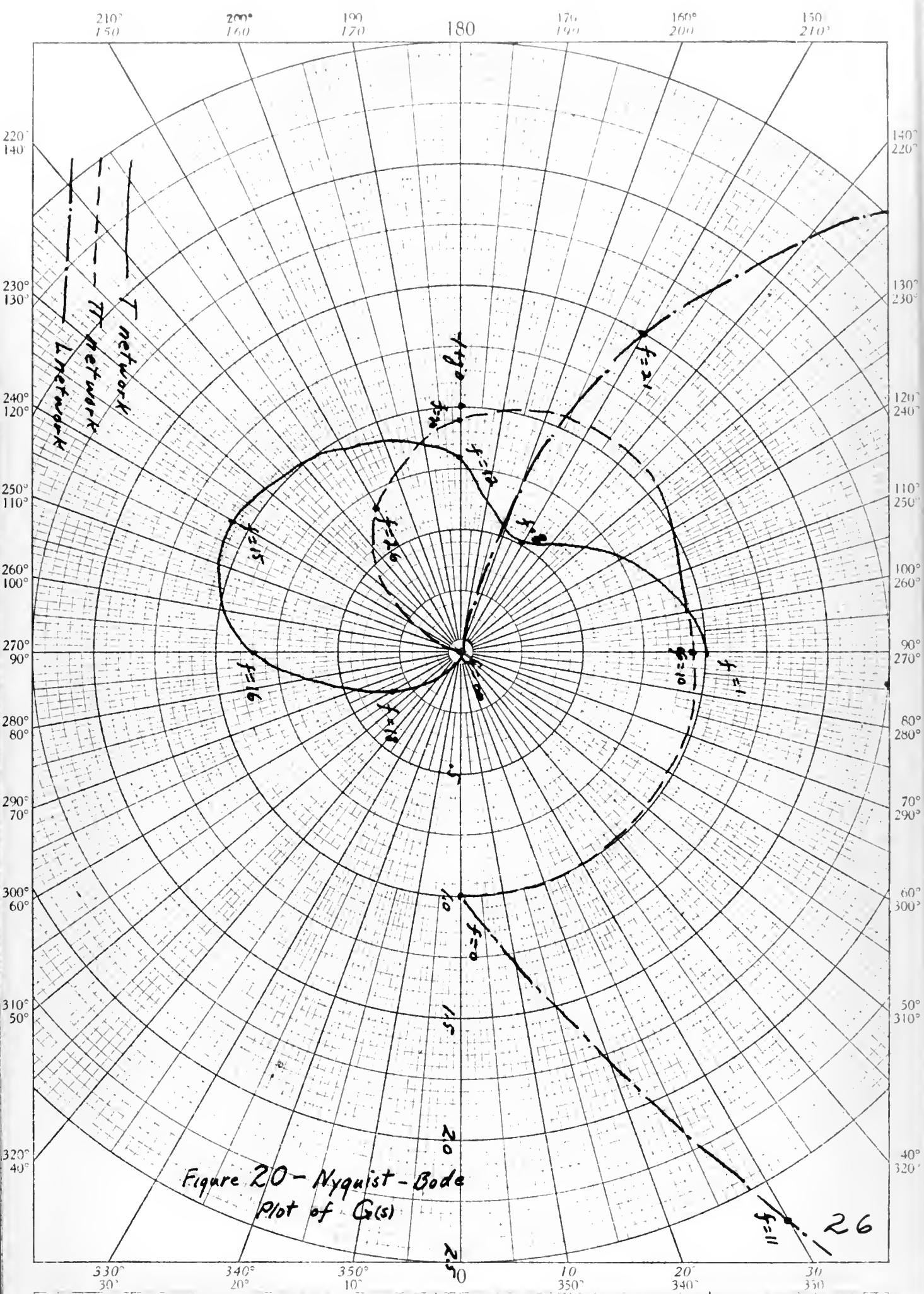
Q arbitrarily set for optimum

Video Compensation









3. The transfer functions of symmetrical pi and T networks are bilateral, while that of the L network is not.

4. Voltages at resonance is the pi circuit divided in inverse proportion to the ratio of input to output capacitance.

5. Voltages divide as the square roots of the input and output impedances.

6. The following equation holds:

$$\rho = \frac{E_{max}}{E_{min}} = \sqrt{Q^2 + 1} = |G(s)|_{max}$$

7. Universal curves are possible for the L network displaying amplitude and phase for various Q's.

8. Universal curves for the pi and T networks are impractical because of the many variables which alter them.

9. Where Q exceeds zero, the impedance match obtained by pi, T and L networks is reliable only at resonance.

[illegible][illegible]

SECRET - 17. 11. 1955 - 18. 01. 1956 - 19. 03. 1956.

• 2001-2002 2003 2004 2005 2006 2007 2008 2009 2010 2011 2012 2013 2014 2015 2016 2017 2018 2019 2020 2021 2022 2023 2024 2025 2026 2027 2028 2029 2030 2031 2032 2033 2034 2035 2036 2037 2038 2039 2040 2041 2042 2043 2044 2045 2046 2047 2048 2049 2050 2051 2052 2053 2054 2055 2056 2057 2058 2059 2060 2061 2062 2063 2064 2065 2066 2067 2068 2069 2070 2071 2072 2073 2074 2075 2076 2077 2078 2079 2080 2081 2082 2083 2084 2085 2086 2087 2088 2089 2090 2091 2092 2093 2094 2095 2096 2097 2098 2099 2100 2101 2102 2103 2104 2105 2106 2107 2108 2109 2110 2111 2112 2113 2114 2115 2116 2117 2118 2119 2120 2121 2122 2123 2124 2125 2126 2127 2128 2129 2130 2131 2132 2133 2134 2135 2136 2137 2138 2139 2140 2141 2142 2143 2144 2145 2146 2147 2148 2149 2150 2151 2152 2153 2154 2155 2156 2157 2158 2159 2160 2161 2162 2163 2164 2165 2166 2167 2168 2169 2170 2171 2172 2173 2174 2175 2176 2177 2178 2179 2180 2181 2182 2183 2184 2185 2186 2187 2188 2189 2190 2191 2192 2193 2194 2195 2196 2197 2198 2199 2200 2201 2202 2203 2204 2205 2206 2207 2208 2209 2210 2211 2212 2213 2214 2215 2216 2217 2218 2219 2220 2221 2222 2223 2224 2225 2226 2227 2228 2229 2230 2231 2232 2233 2234 2235 2236 2237 2238 2239 2240 2241 2242 2243 2244 2245 2246 2247 2248 2249 2250 2251 2252 2253 2254 2255 2256 2257 2258 2259 2260 2261 2262 2263 2264 2265 2266 2267 2268 2269 2270 2271 2272 2273 2274 2275 2276 2277 2278 2279 2280 2281 2282 2283 2284 2285 2286 2287 2288 2289 2290 2291 2292 2293 2294 2295 2296 2297 2298 2299 2300 2301 2302 2303 2304 2305 2306 2307 2308 2309 2310 2311 2312 2313 2314 2315 2316 2317 2318 2319 2320 2321 2322 2323 2324 2325 2326 2327 2328 2329 2330 2331 2332 2333 2334 2335 2336 2337 2338 2339 2340 2341 2342 2343 2344 2345 2346 2347 2348 2349 2350 2351 2352 2353 2354 2355 2356 2357 2358 2359 2360 2361 2362 2363 2364 2365 2366 2367 2368 2369 2370 2371 2372 2373 2374 2375 2376 2377 2378 2379 2380 2381 2382 2383 2384 2385 2386 2387 2388 2389 2390 2391 2392 2393 2394 2395 2396 2397 2398 2399 2400 2401 2402 2403 2404 2405 2406 2407 2408 2409 2410 2411 2412 2413 2414 2415 2416 2417 2418 2419 2420 2421 2422 2423 2424 2425 2426 2427 2428 2429 2430 2431 2432 2433 2434 2435 2436 2437 2438 2439 2440 2441 2442 2443 2444 2445 2446 2447 2448 2449 2450 2451 2452 2453 2454 2455 2456 2457 2458 2459 2460 2461 2462 2463 2464 2465 2466 2467 2468 2469 2470 2471 2472 2473 2474 2475 2476 2477 2478 2479 2480 2481 2482 2483 2484 2485 2486 2487 2488 2489 2490 2491 2492 2493 2494 2495 2496 2497 2498 2499 2500 2501 2502 2503 2504 2505 2506 2507 2508 2509 2510 2511 2512 2513 2514 2515 2516 2517 2518 2519 2520 2521 2522 2523 2524 2525 2526 2527 2528 2529 2530 2531 2532 2533 2534 2535 2536 2537 2538 2539 2540 2541 2542 2543 2544 2545 2546 2547 2548 2549 2550 2551 2552 2553 2554 2555 2556 2557 2558 2559 2560 2561 2562 2563 2564 2565 2566 2567 2568 2569 2570 2571 2572 2573 2574 2575 2576 2577 2578 2579 2580 2581 2582 2583 2584 2585 2586 2587 2588 2589 2590 2591 2592 2593 2594 2595 2596 2597 2598 2599 2600 2601 2602 2603 2604 2605 2606 2607 2608 2609 2610 2611 2612 2613 2614 2615 2616 2617 2618 2619 2620 2621 2622 2623 2624 2625 2626 2627 2628 2629 2630 2631 2632 2633 2634 2635 2636 2637 2638 2639 2640 2641 2642 2643 2644 2645 2646 2647 2648 2649 2650 2651 2652 2653 2654 2655 2656 2657 2658 2659 2660 2661 2662 2663 2664 2665 2666 2667 2668 2669 2670 2671 2672 2673 2674 2675 2676 2677 2678 2679 2680 2681 2682 2683 2684 2685 2686 2687 2688 2689 2690 2691 2692 2693 2694 2695 2696 2697 2698 2699 2700 2701 2702 2703 2704 2705 2706 2707 2708 2709 2710 2711 2712 2713 2714 2715 2716 2717 2718 2719 2720 2721 2722 2723 2724 2725 2726 2727 2728 2729 2730 2731 2732 2733 2734 2735 2736 2737 2738 2739 2740 2741 2742 2743 2744 2745 2746 2747 2748 2749 2750 2751 2752 2753 2754 2755 2756 2757 2758 2759 2760 2761 2762 2763 2764 2765 2766 2767 2768 2769 2770 2771 2772 2773 2774 2775 2776 2777 2778 2779 2780 2781 2782 2783 2784 2785 2786 2787 2788 2789 2790 2791 2792 2793 2794 2795 2796 2797 2798 2799 2800 2801 2802 2803 2804 2805 2806 2807 2808 2809 2810 2811 2812 2813 2814 2815 2816 2817 2818 28

تاریخ: ۱۳۹۸/۰۵/۰۵

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...I told them that I was not at home, or, leaving Israel.

2. Strategy to occupy the institutions

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CHAPTER IV

DISTRIBUTED AND CASCADED AMPLIFIERS

(56)

One type of broadband amplifier patented by Percival in England and developed in this country by Spencer, Hewlett and Kennedy is the distributed amplifier which makes use of artificial transmission line sections to reduce the effect of tube capacity, provide a constant load impedance to the tubes, and a relatively constant gain over a broadband of frequencies up to about 220 mc. Using special tubes, Melpar Incorporated has developed an amplifier of this type usable to about 1,000 mc.

The limit in gain for amplifiers of this type is approximately 20 db, and while they are used in special applications such as millimicrosecond oscillography, for television they have largely been superseded by a series of cascaded triodes with either pi or M derived coupling sections. Triode boosters for television using 4 dual triodes give a 24 db gain with a noise figure of 12 db over the television bands. Pentode distributed amplifiers use 8 pentodes give 22 db of gain, with a 20 db noise figure over the same band.

(48)(49)

Figure (21) shows the basic Spencer-Kennedy circuit. In designing this circuit, the L is usually made a continuous coil on a long rod. The tube elements furnish the capacity - except that some slight additional capacity is needed in the plate circuits to make the grid and plate artificial lines identical.

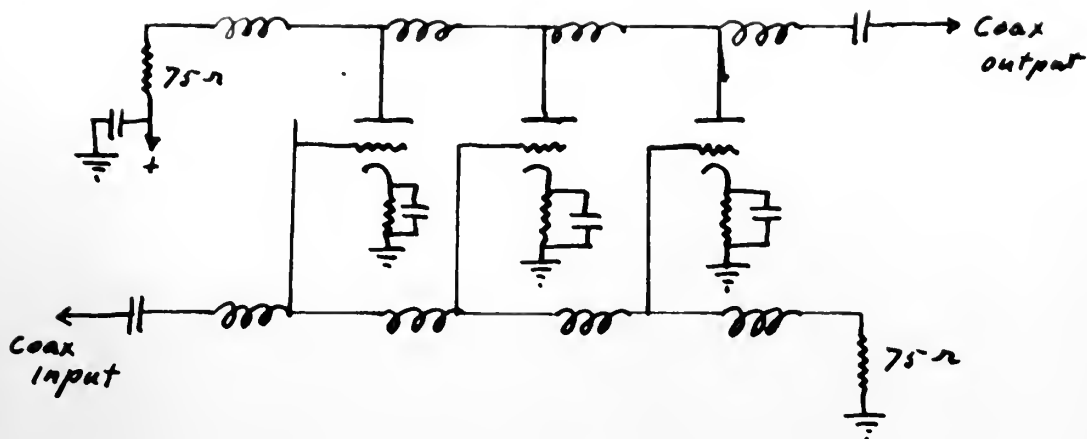
There are more involved concepts to explain the circuit, but the simplest is that the tubes are all in parallel, and the only function

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 seventh of these is the fact that the
 eighth of these is the fact that the
 ninth of these is the fact that the
 tenth of these is the fact that the

Figure (1) shows the basic Spectroscopy principle. This circuit, the most commonly used, is shown in Figure (1). The basic principle of Spectroscopy is shown in Figure (1). The basic principle of Spectroscopy is shown in Figure (1). The basic principle of Spectroscopy is shown in Figure (1).

and the following is a list of addresses of persons who are interested in the work of the Commission.



For Pentodes

$$G_m' = n \times G_m$$

$$R_L = \frac{1}{2} Z_0 = 37.5 \Omega$$

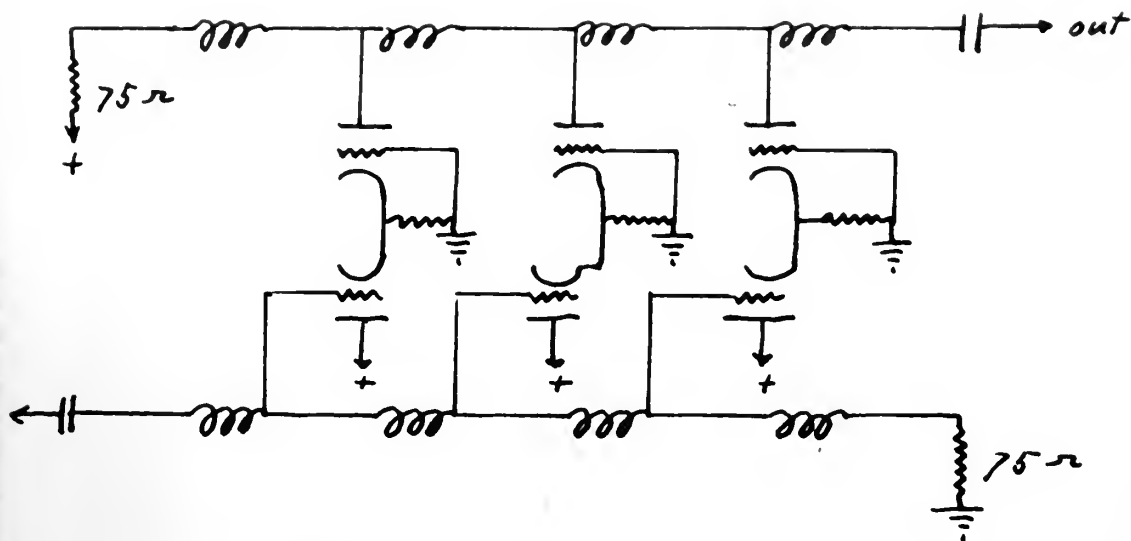
Assume 8 - 6AK5's

$$G_m = 5,000 \mu\text{mhos}$$

$$G_m' = 40,000 \mu\text{mhos}$$

$$\text{Gain} = R_L' G_m' = 37.5 \times 40 \times 10^{-3} = 15 = 23 \text{ db}$$

Figure 21 - The Distributed Amplifier



$$\text{Gain} = \frac{K \mu R_L'}{R_p' + R_L'}$$

$K =$ a loss factor due to paraphase coupling
 $\approx .7$

Assume $\mu = 40$

$$R_p = 5,000 \Omega$$

8 Tubes

$$R_p' = 625$$

$$R_L = 37.5$$

$$G = \frac{40 \times 37.5 \times .7}{625 + 37.5} = 1.6 = 4 \text{ db}$$

Figure 22 - The Paraphase Amplifier



of the line is to isolate tube capacities and give a constant load over the pass band.

Because of the high equivalent noise resistance of the pentodes, (about four times that of a VIF triode), triodes are preferable, but the great difficulty in their use is the need for neutralization. Melpar has devised a circuit using special pencil triodes which is usable to 1000 mc, and does not require neutralization. This circuit is a cathode follower driving a grounded grid section. Melpar prefers to call it a paraphase circuit.

Figure (22) shows the all triode circuit.

Tubes of the 6J6 type are usable to 200 mc. For circuits up to 1000 mc., pencil triodes with a special isolated cathodes are used. In all cases, the filaments must be floating at R. F. by effective R. F. chokes. For the 1000 mc. unit described at the 1953 IRE convention, the inductances were merely straps connecting the tube sockets.

The gain of such a circuit is given in figure (22). It can easily be seen that special tubes with low R_p and high μ are required if appreciable gain is to be realized.

Because of the economy factor in designing an item for home use, as well as for reasons of better performance, present day boosters use pi couplings or M dervied sections between cascaded triode sections.

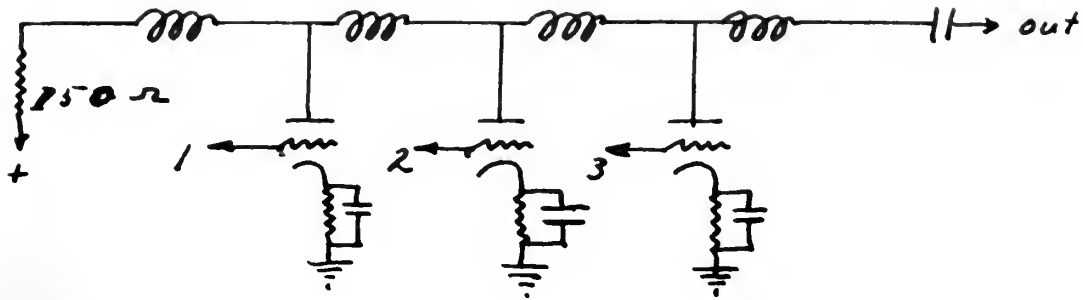
Distributed amplifiers are not limited to amplifying. They are also useful for mixing and distributing signals. Figure (23) shows a three channel mixer with separate grid inputs, but with all plates forming elements of the line. Pi or L couplers may be used at the grids

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• 2012年10月10日，中国首条高速铁路——京沪高速铁路全线通车。

100



For Triodes

$$\text{Gain} = \frac{\mu R_L}{R_L + R_p} = \frac{40 \times 75}{75 + 5,000} = .6$$

For Pentodes

$$\text{Gain} = G_m R_L = 5,000 \times 10^{-6} \times 75 = .375$$

Figure 23 - The Distributed Mixer

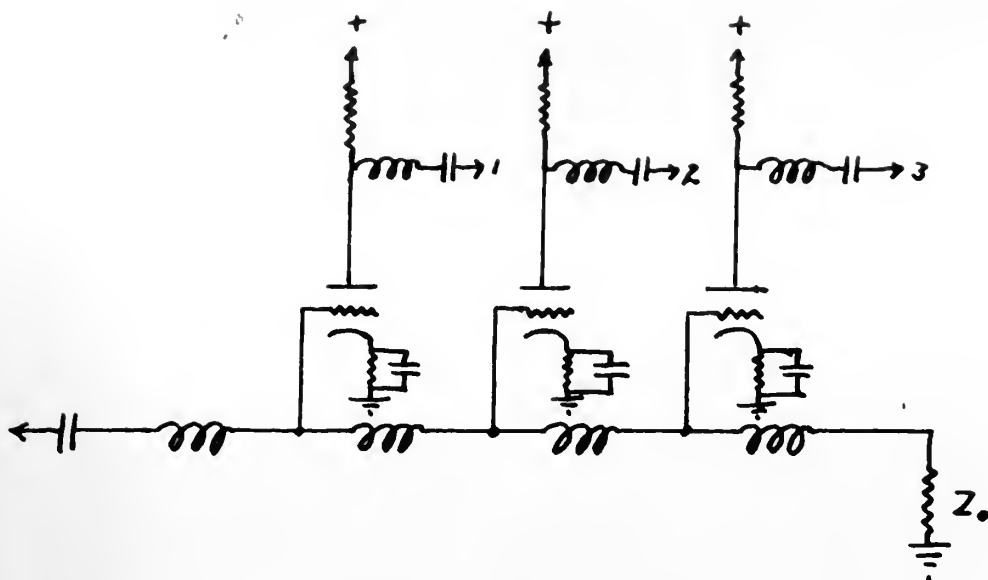


for impedance matching and voltage step up. In this case the Q of the input match is limited by the bandpass required of the input, while the output line usually is made a very low Q or flat.

There is some loss in mixing. Triodes are suitable without neutralizing because the Miller effect is negligible when the gain is less than 1. For television circuits using a 6BQ7 or 6BK7, the line to pass all 13 VHF channels has a maximum $Z_0 = 150$ ohms, thus $R_L = 75$ ohms. Triode mixers are to be preferred because of gain and noise figure. Usually they are used in pairs to provide a 300 ohm balanced input and output.

The line may be used in the grid circuit to feed several isolated outlets as in figure (24). Here an artificial line feeds three separate outlets. If the outlet is to utilize only part of the frequencies carried, the L or pi matched plate circuit may have a Q greater than 0. If it is to be broadband, Q is set at 0 or 1. R is large so that its effect on Q is negligible compared to the reflected load of the output. Pentodes may be used singly, or triodes in pairs. The distribution unit has a slight gain.

Triodes have a greater feedback factor so that they offer less isolation to the outputs. For this reason, pentodes are preferred in commercial units. Using pi couplings with a Q of 1 tuned to about channel 11, the pentode units can drive an effective R_L large enough to give a slight gain over most of the spectrum. Because they are single ended however, matching transformers are necessary to feed 300 ohm lines, and the overall effect is a gain of about 1/1. Again, because the line



For Triodes $\text{Gain} = \frac{\mu R_L}{R_L + R_p} = \frac{40 \times 150}{150 + 5,000} = 1.16$

For Pentodes

$$\text{Gain} = G_m R_L = 5,000 \times 10^{-6} \times 150 = .75$$

Figure 24 - Signal Distribution System

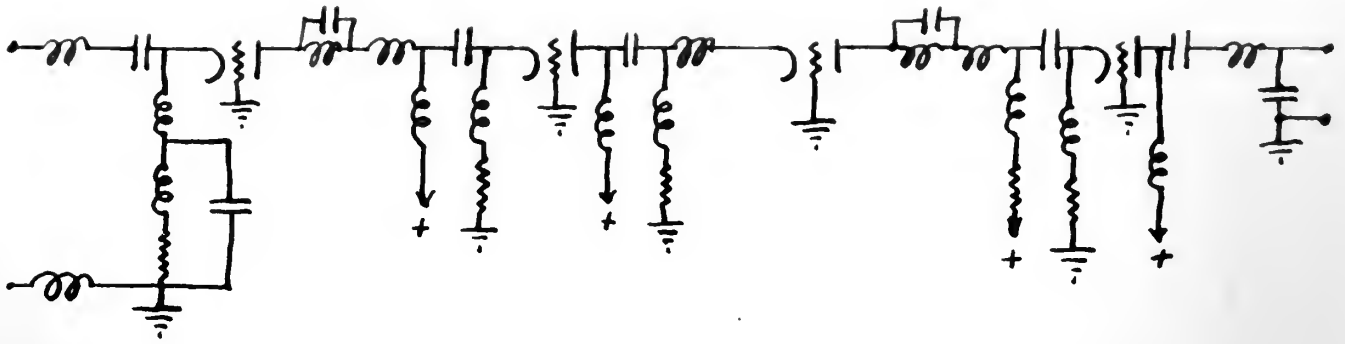


Figure 25 -
Cascaded Triode R.F. Amplifier

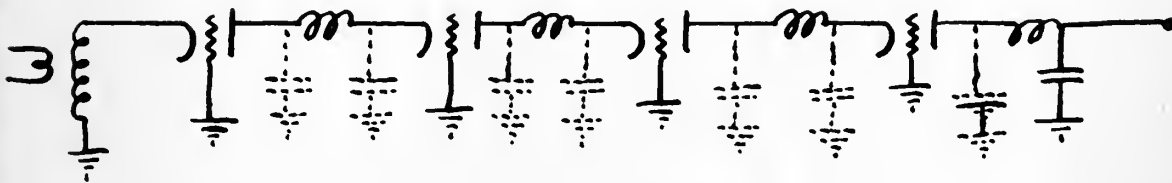
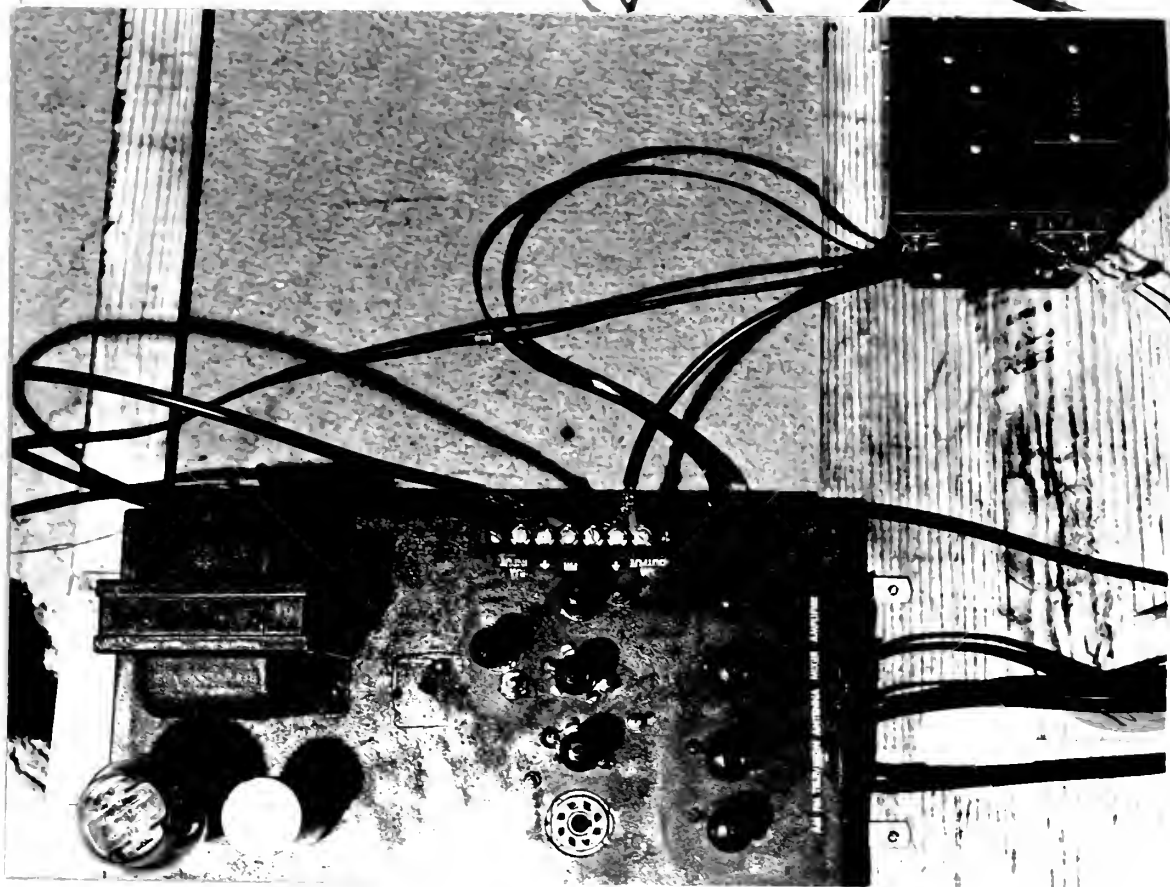
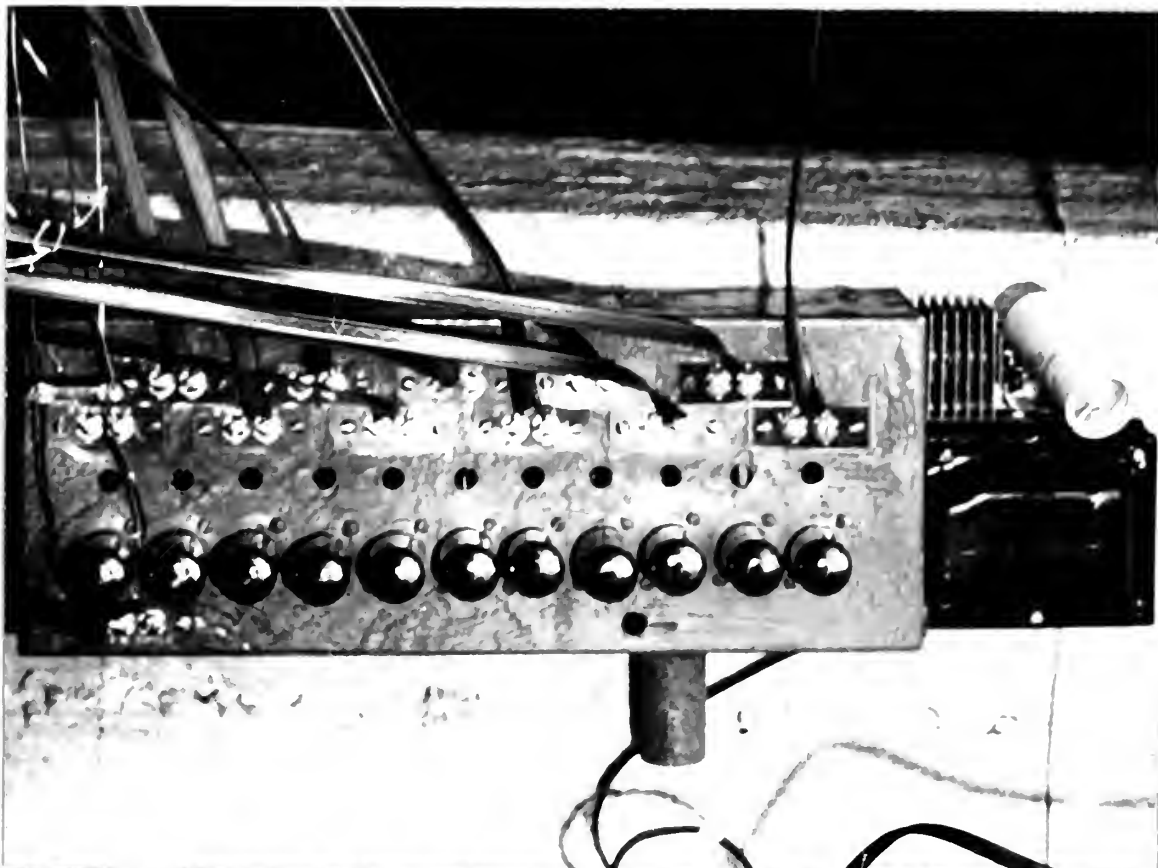


Figure 25(a)
Simplified Circuit



is approximately 150 ohm to cover the VHF Television range, triodes are used in pairs to feed from and distribute to 300 ohm balanced lines.

Figure (26) shows photographs of an 11 outlet distributed line distribution unit built at the U. S. Naval Postgraduate School for use in the B. O. Q. Figure (27) shows a wide band booster under a 4 outlet distribution unit built at the U. S. Naval Postgraduate School for use in the B. O. Q.

Figure (25) shows the IT-102A auto-booster which gives 19 db of gain with a 9 db noise figure.

While the similarity of this unit to a group of pi couplings is hard to see, the circuit actually is a development of the one in figure (25(b)) which ignores voltage feeds and isolating RF chokes. Series Peaking circuits are added to flatten response, and the R. F. chokes are made resonant at some mid-frequency to boost gain. Using grounded grid tubes throughout reduces the noise to an acceptable level. The Blonder-Tongue Company makes a similar unit in which only the first tube is grounded grid. The noise figure is almost doubled.

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CHAPTER V

INTERSTAGE COUPLING

The use of pi sections as impedance matching sections is not a recent innovation. It is, in fact, almost as old as the art itself. Several of the earlier circuits are shown in figure (28).

In post World War II television circuits T and pi networks are used to couple I. F. stages, video stages, and more recently R. F. stages in the form of cascade or driven grounded grid R. F. amplifiers and boosters.

Perhaps the best example of how the pi coupling has developed is the driven grounded grid. Figure (29(a)) shows the circuit as originally used by television manufacturers.

Figure (29(a)) shows the basic driven grounded grid circuit using a special dual triode such as the 6BQ7, 6BK7 or 6BZ7. An improvement is shown in 29(b). The inductance L_1 serves two purposes, first, to resonate with the plate and cathode capacitances and thus match impedances; second, to cause a 180° phase shift between plate and cathode, thus giving almost perfect neutralization because input and output are 180° out of phase and any feedback is negative, this results in 1 or 2 db improvement in noise figure. Figure (29(c)) shows a further refinement for higher gain with good neutralization. The tapped inductance L_1 matches a low impedance cathode (150 ohms) to a plate at about 1500 ohms with a modified pi match. With the plate driving the cathode directly, the first stage has a gain of approxi-

[illegible]

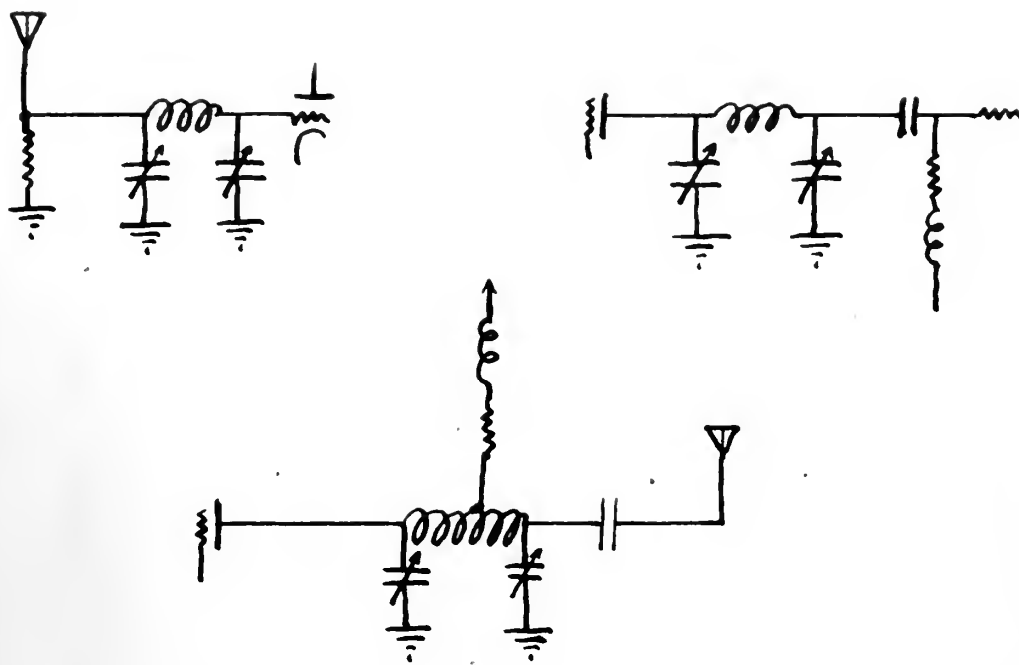


Figure 28 - Matching Sections (π)

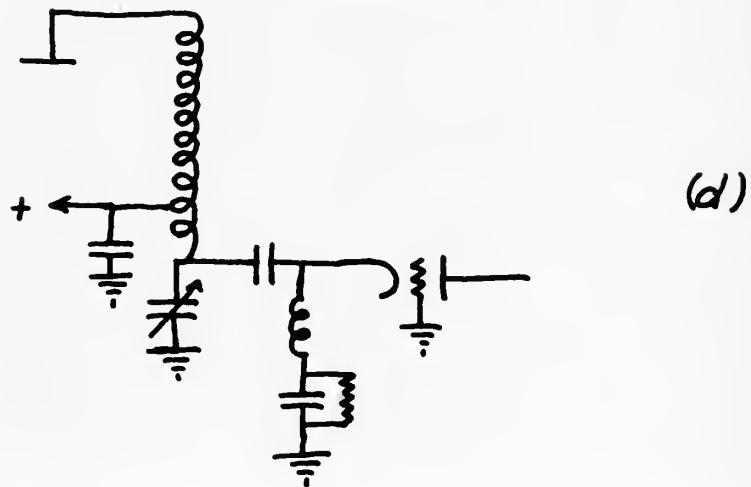
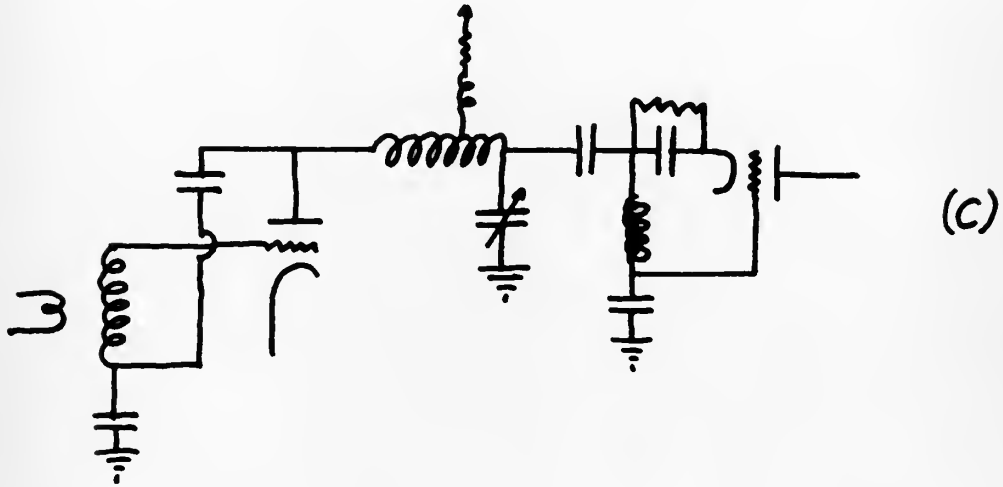
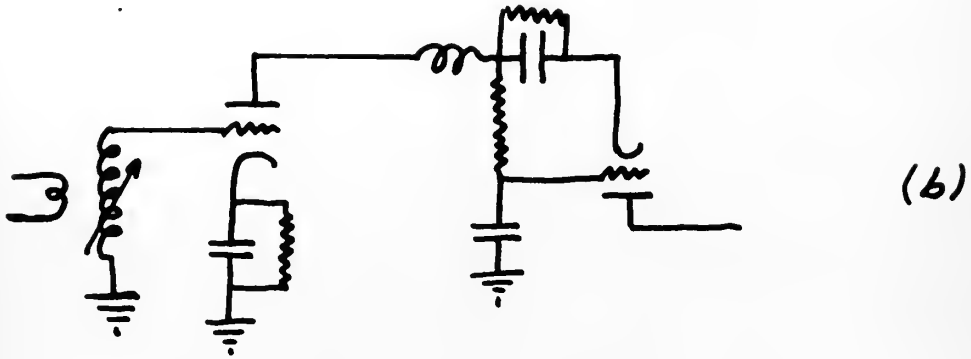
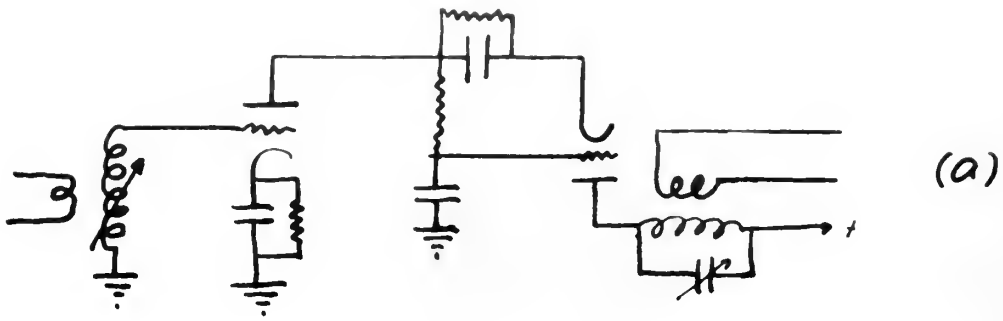


Figure 29 - Cascode Circuits

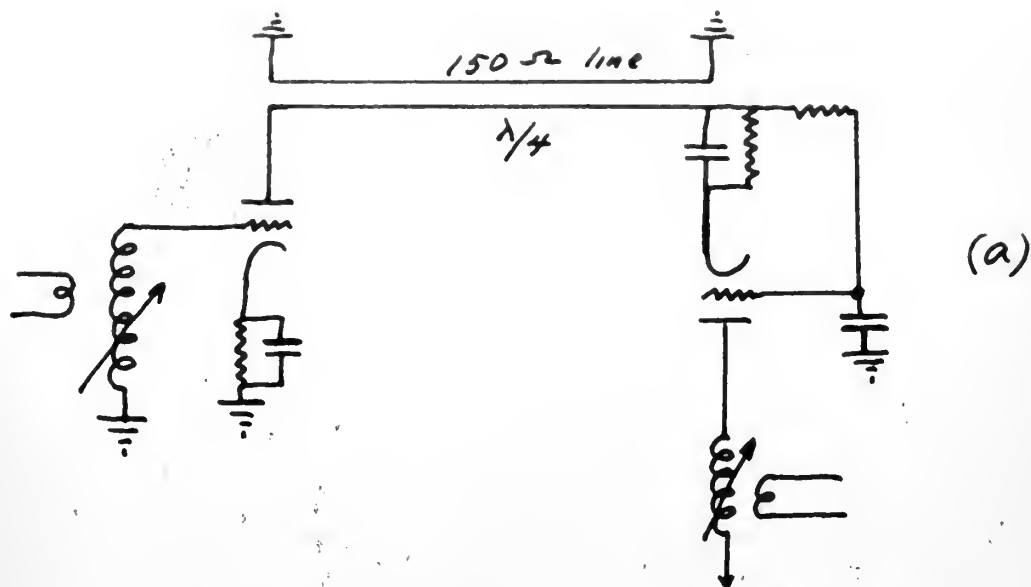
mately 1. Using the auto transformer L_1 the gain is $\sqrt{\frac{1500}{150}} =$
 3.2 or approximately 10 db. The output phase is still negative so that
 good neutralization is achieved. L_2 is used to resonate with the
 cathode to obtain a pure resistive load to the output of L_1 .
 (52)

An interesting variation of this circuit used by Philco uses an
 actual transmission line section of 150 ohms twinlead. This circuit is
 shown in figure (30). The plate sees the end of a transmission line
 loaded with its Z_0 , hence the load on the plate is 150 ohms. For any
 feedback signal however, the plate is essentially an open circuit, and
 the cathode sees a short. In this way better neutralization is achieved.

The pi section may be used in IF stages to obtain a relatively flat
 response through the I.F. region, but rejecting the fundamental R. F.
 and oscillator frequencies. The circuit shown in figure (30(b)) is
 (52)
 used by Admiral to couple the mixer pentode to the first I. F. by load-
 ing the plate with a small additional condenser is obtained, and the
 resulting pi has a Q of 2 or 3 with a sharp cutoff at about 30 mc.

Artificial line sections are not frequently used as I.F. inter-
 stage couplings, most manufacturers preferring stagger tuned circuits
 with traps or M derived filter sections.

[illegible]



$$Z_o = R_L = \frac{1}{G_m} = 150 \Omega$$

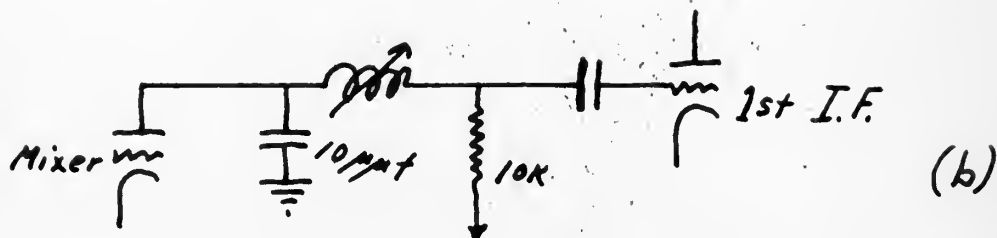


Figure 30
Tuner And Mixer Coupler Circuits Used By
Admiral and Philco



CHAPTER VI

THE VIDEO AMPLIFIER

Compensation of a video amplifier to get flat response may be treated as an extension of the basic line theory. Figure (31(a)) shows the video amplifier compensation as normally drawn. Figure (31(b)) shows the compensation redrawn as a ladder network of two L sections.

The capacitors C_1 and C_2 are tube output and input capacitors. It is interesting to note that the feed point may be shifted to the center of the ladder if plate capacity is smaller than grid capacity as in figure (31(c)).

The simple case of series peaking is merely one of adjusting a pi network for maximum flat response. The plate supply resistor is placed on the side where C is smallest, in this way the circuit Q is controlled at high frequencies with maximum gain at lower frequencies. See figure (31(d)).

The method of design is to find the 3 db down point of the uncompensated amplifier, i.e., the frequency at which total capacitive reactance equals the plate load resistance, then pick an L to resonate at 1.4 times this frequency at a Q of 1 with R_1 as the shunting Q limiting factor and with C_1 and C_2 in series. Obviously this requires some juggling of values to obtain, but when once adjusted, a very good response curve is obtained.

It should be stated here that while the series peaking and combination peaking circuits give higher gain, they are in disrepute in high quality systems because their sharp cutoff tends to cause ringing by

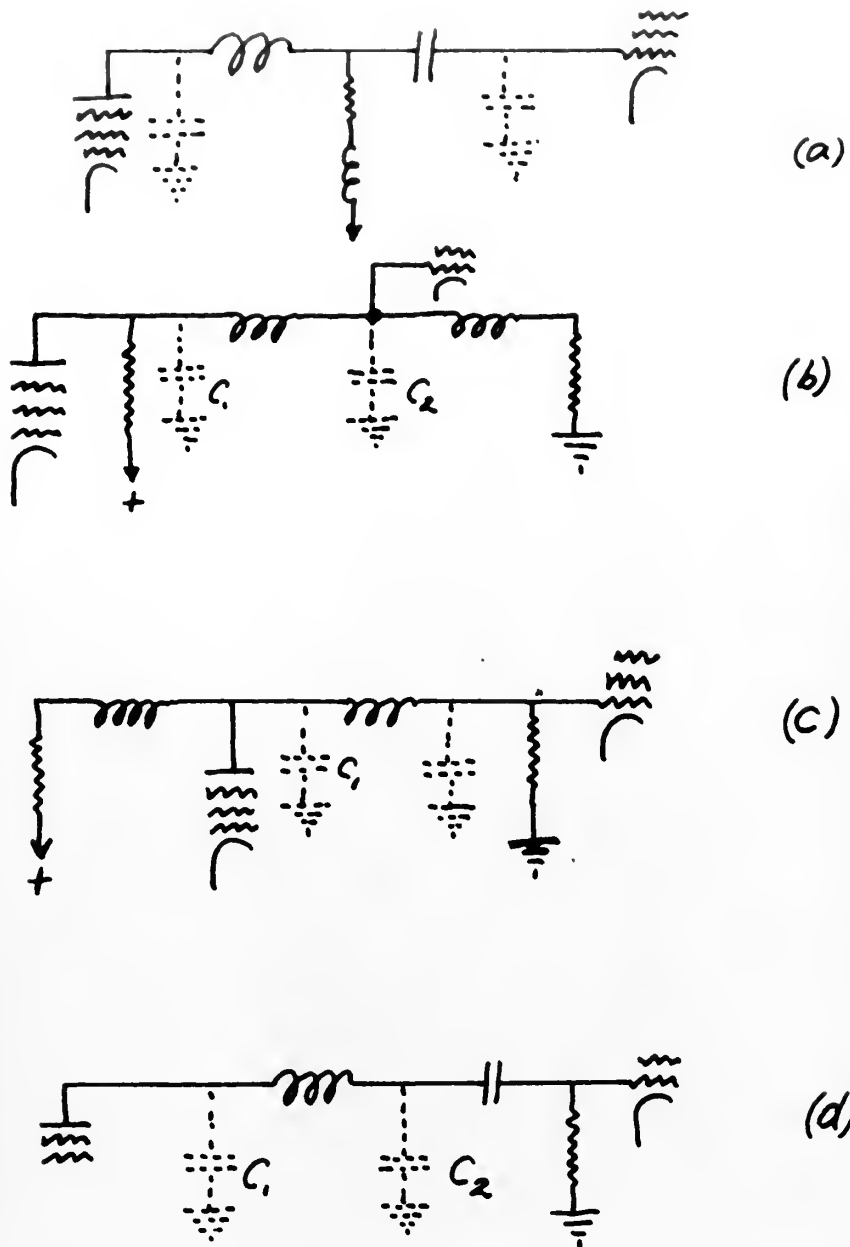
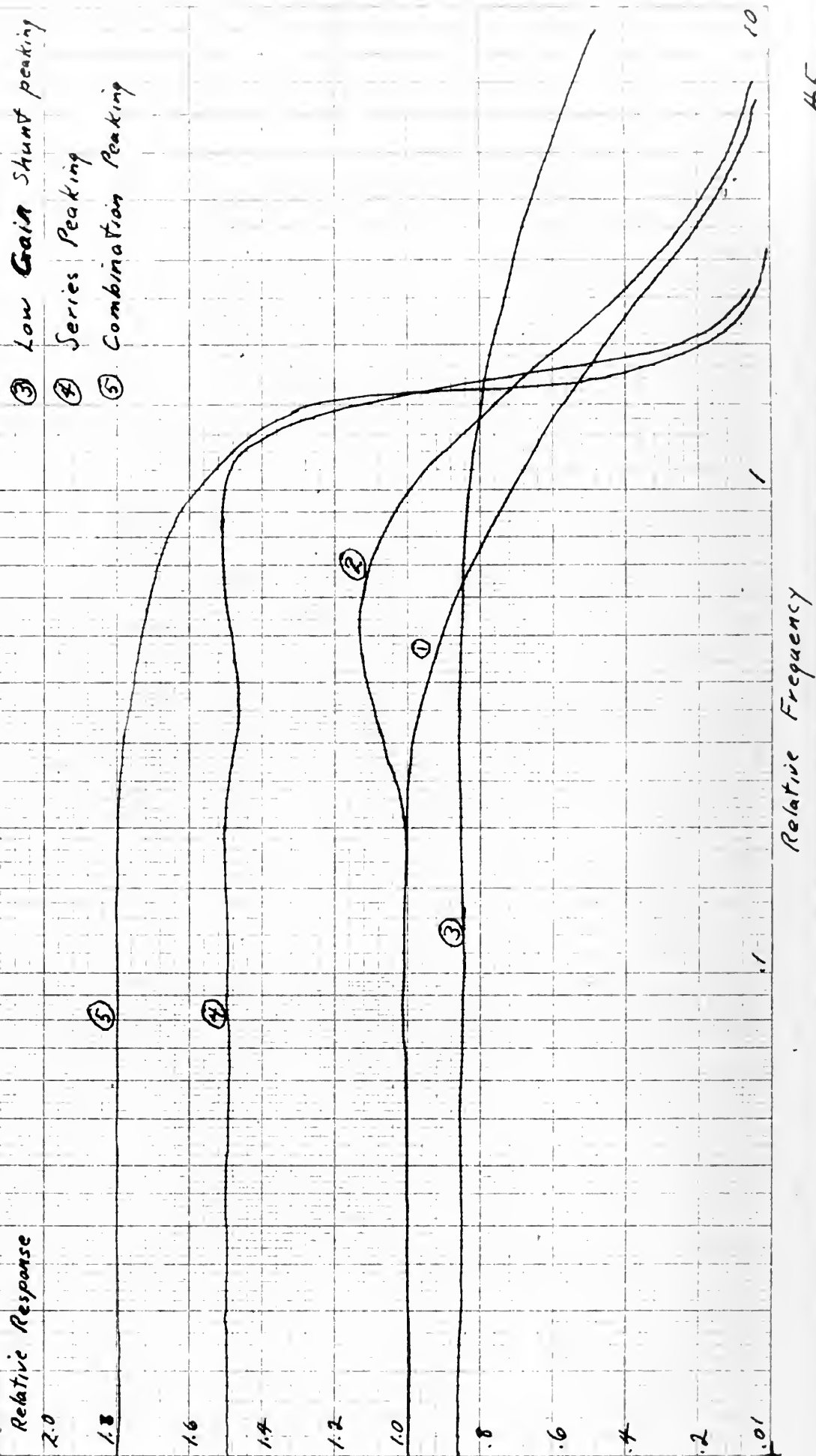


Figure 31 - Video Coupling Circuits

Figure 3R - Video Response Curves With Various Coupling Methods.

- ① uncompensated
- ② Shunt Peaking
- ③ Low Gain Shunt peaking
- ④ Series Peaking
- ⑤ Combination Peaking



(37)(39)
transients and consequent distortion of the picture. Combination and series peaking are little used in transmitters, and find their chief application in receivers where the cutting off has already been done by transmitting duplexers and receiver I. F. stages. Shunt peaking is to (37)(39) be preferred in television circuits in spite of the lower gains achieved.

Methods for calculating the optimum values for these video peaking (57) systems have been worked out, and the rules for determining optimum (53)(54)(55) values are found in nearly all texts, so that with the possible exception of the pi match or series peaking, little or no advantage is obtained by using the methods set forth in this thesis for video compensation. Figure (32) shows the relative responses of the various methods of video compensation.

[illegible]

CHAPTER VII

EXPERIMENTAL CIRCUITS

The use of pi sections presents several interesting possibilities in R. F. circuits. One of these is the possibility of achieving an extremely low noise figure for I. F. stages coupled to a balanced mixer, or for television R. F. stages coupled direct to balanced inputs. Figure (33) shows the basic circuit of the low noise input devised by the author.

In determining noise figure, the actual circuit resistance is compared with a theoretically perfect tube so that the actual input resistance divided by the ideal input resistance yields have equation (21). Figure (34) shows the equivalent circuits of the actual and ideal cases. (33)(34)(43)
The tube is considered ideal.

If R_1 is made large, the effect of noise resistance is decreased, i.e., if we use a high Q circuit at the input, the noise figure will decrease with increasing Q. A point is soon reached, however, in television circuits where the bandwidth becomes too narrow, or tube input conductance limits Q by its shunt resistance effect.

The circuit of figure (33) uses 2 pi coupling units at the input which are higher than normal Q for the bandpass required and stagger tuned. Looking in at the low impedance end, the Q is very low and the feed system sees a balanced input, but at the grid side the Q is high and the input is thus frequency selective. As an example of what might be done with an input of this type, assume a signal is desired at 69 mc (channel 4) with a 6 mc bandpass. To do this with a conventional input,

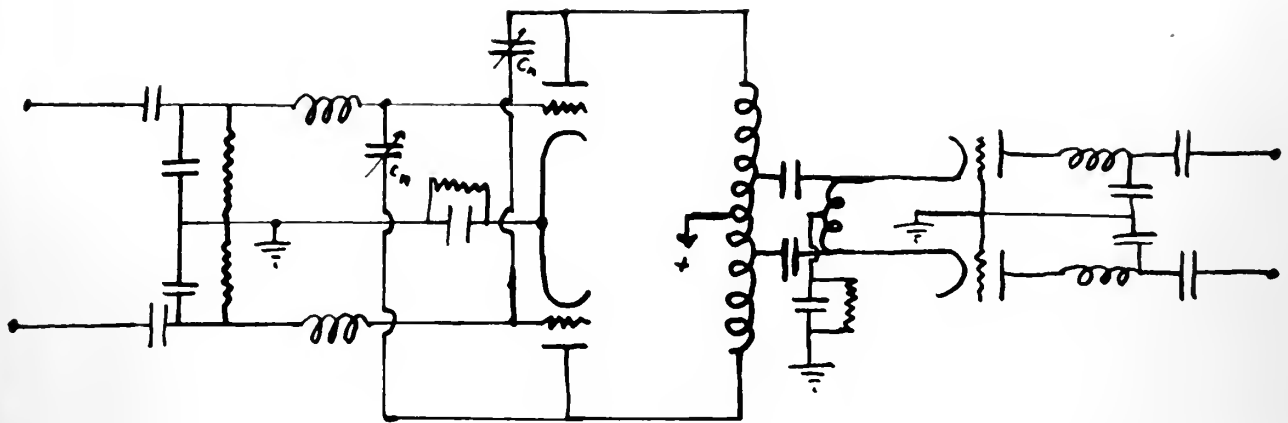
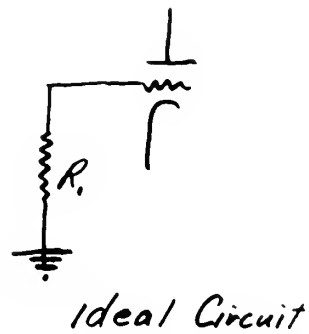
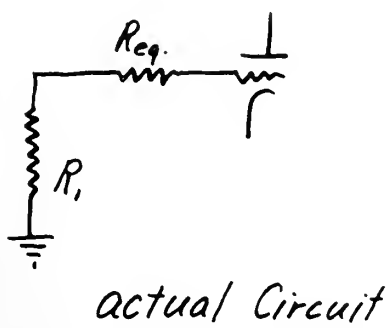


Figure 33 - A Low Noise R.F. Amplifier



$$\begin{aligned}
 E_n &= K \sqrt{R_{eq}} \\
 &= \sqrt{4kT\Delta f} \sqrt{R_{eq}} \quad (20)
 \end{aligned}$$

$$n.f. = \frac{R_i + R_{eq}}{R_i} \quad (21)$$

Figure 34 Actual and Equivalent Circuits For Determining Noise Figure.

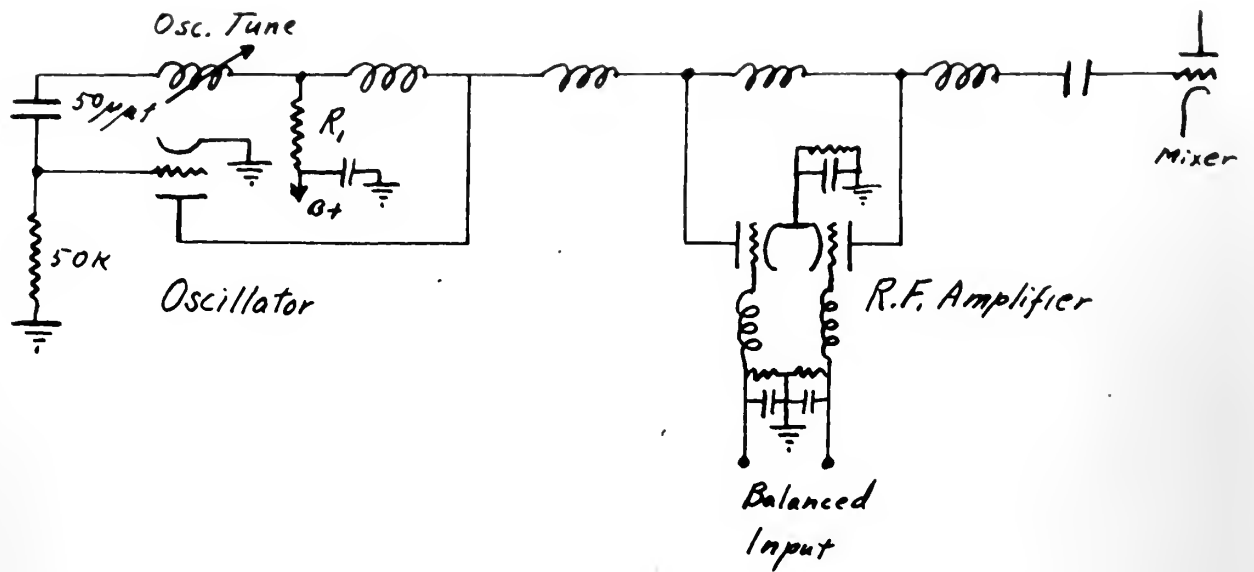


Figure 35-A Distributed Line Converter.

the Q cannot exceed 10 or 11. Using stagger tuned pi units, one tuned to 67.5 and the other to 70.5, the Q can be made equal to 20 and a 6 mc bandpass, with a flatter top characteristic can be maintained. In so doing, the resistance R_1 is raised and the effect of R_{eq} is decreased. An improvement in noise figure up to 3 db is possible.

A unit to test this circuit was built by the author and tested in January of 1953 using 2-6BK7 tubes. Making measurements with a Kay Mega-node a noise figure of 3 db was obtained.

Another interesting possibility is the use of a long line section for oscillator tank, mixer, and R. F. stage. Figure (35) shows the circuit devised by the author.

In figure (35) the inductances and tube elements form a line which has a relatively high Q at the oscillator, and R. F. The mixer grid being a mismatch focal point equivalent to an open ended line. Standing waves are created by R_1 which is smaller than Z_0 .

Such a circuit has little to recommend it at lower frequencies, but at the very high or U.H.F. regions the use of simple loops or hair-pin inductances with slide tuners can result in a very desirable unit, chiefly because no inductive coupling is required and many components such as R. F. chokes and by-pass condensers are unnecessary.

The oscillator will oscillate at the frequency at which its inter-electrode inductance gives a 180° phase shift. The Q of the oscillator tank is determined by the tuning of the plate sections and R_2 . The mixer has maximum gain at the frequency at which its interplate inductance has a 180° phase shift. Both sections pass their outputs down the artificial

Relative Response

Figure 36 - Test Data of The Circuit
Given in Figure 35

— R.F. Voltage
At Mixer - E_m - Constant
--- Oscillator Voltage -
L variable to tune -
Measured at Mixer.

frequency

52

100

10

1

0.2

0.4

0.6

0.8

1.0

1.2

1.4

1.6

1.8

2.0

2.2

2.4

2.6

2.8

3.0

3.2

3.4

3.6

3.8

4.0

4.2

4.4

4.6

4.8

5.0

5.2

5.4

5.6

5.8

6.0

6.2

6.4

6.6

6.8

7.0

7.2

7.4

7.6

7.8

8.0

8.2

8.4

8.6

8.8

9.0

9.2

9.4

9.6

9.8

10.0

line so that they tend to reflect from end to end causing a standing wave at the mixer.

The circuit of figure (35) has been built and tested. Results of test data and curves are given here in figure (36). In order to get the most reliable data in the easiest way, a technique which might be called frequency scaling is used. Large inductances and capacities are employed so that the system which would normally be operating in megacycles can be tested in the kilocycle range and such factors as amplitude and phase can be observed on an oscilloscope.

Several other circuits employing line sections are given in the Appendix without explanation or comment.

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...the ... of the ... to ... the ... of the ...

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APPENDIX A

Examples of Applications of Figure 3

I. Plate to Grid Coupling

A given plate circuit has $10 \mu\text{f}$ of capacity which is 1000 ohms X_C at the operating frequency. The following grid has also $10 \mu\text{f}$ of capacity. The bandpass requirements are such that a $Q = 10$ will serve. Since a condenser input and output is present, pi coupling is indicated. The $Q = 10$ is large enough that X_L can be made equal to X_C , and R_2 will equal $10 X_C$ 10,000 ohms. R_1 will equal 100 ohms. Referring to the pi equivalent in figure (4), R_1 is an internal impedance at the center and is not an actual resistor. R_2 must be an actual resistor at one end, to supply B \nearrow to the plate. Since there are 2L sections to make the pi, the L chosen will be twice the value indicated above, or X_L 2000 ohms. The final design is shown in figure (5). This is an actual mixer to I. F. coupling used by Admiral.

II. Line to Grid Coupling, or plate to line

It is desired to connect a 6AK5 grid to a 75 ohms line. The input capacity is $11 \mu\text{f}$. The frequency range must be as broad as possible, preferably reaching 200 mc. Referring to reactance charts figure (6) we find $11 \mu\text{f}$ has a reactance of 150 ohms at 100 mc, or 75 ohms at 200 mc. We can make a one to one match with Q equal zero at 200 mc. Referring to figure (3), we note the ratio of X_C to X_L for Q equal 0 is infinite. An arbitrary decision must therefore be made. Since the loading resistance is the line, the phase can be kept near zero shift if we make

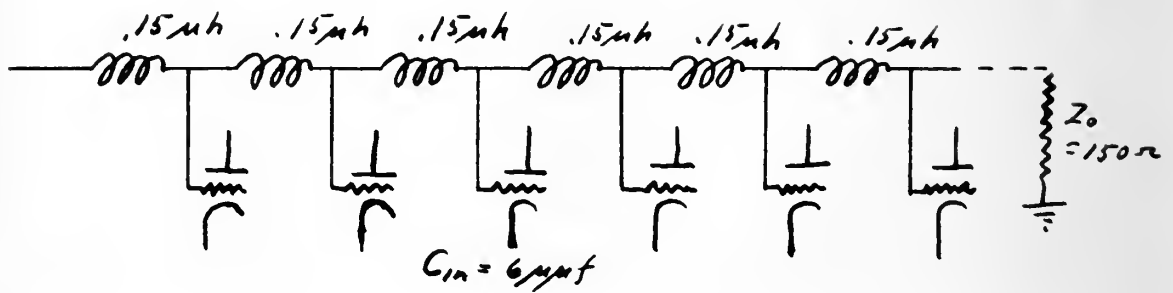
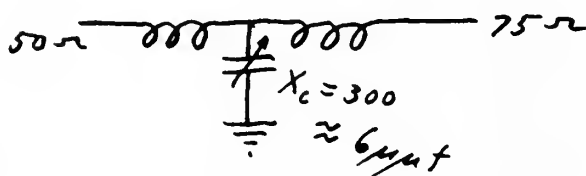
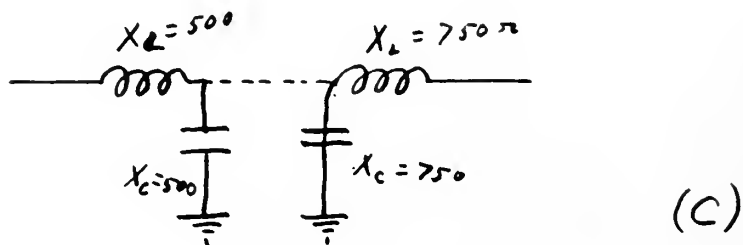
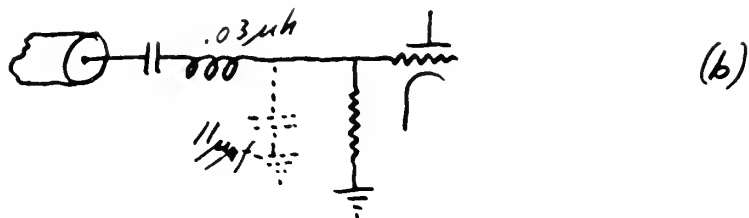
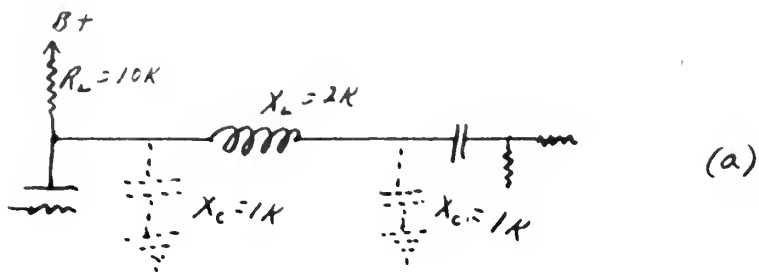


Figure 5 - Examples of Impedance Matches.

$X_C = X_L$ at 1.4 times 100 mc or 280 mc. At this frequency $X_C = 53$ ohms, so X_L is picked as 53 ohms at 280 mc, or 38 ohms at 200 mc, or 19 ohms at 100 mc, which corresponds to .03 microhenry. The same approach can be used for plate to line coupling. See figure 5(b).

III. Line to Line Coupling

It is desired to match a 50 ohms line to a 75 ohms line with a frequency selective match. An overall Q of 10 at 100 mc. is desired. A T match is indicated. As was shown in Chapter II, the Q will stay 10 if two Q = 10 sections are linked. If Q = 10, $R_1 = 50$ ohms, $X_C = 500$ ohms. Since Q is large X_L will also equal 500 ohms. R_2 , which is an internal impedance and not an actual resistor or load is 5000 ohms. These values can easily be picked off the chart in figure 3. Since the T contains two L sections joined at R_2 , the actual value of C would be the two values paralleled. Such a circuit would have to be tuned for best results. See figure 5(c).

IV. Line to Multiplate Stage Coupling

Ten grids must be paralleled to be driven from a 150 ohms line. the 12AV7 which has a total input capacity without Miller effect at 6 is to be used. Ten grids at 6 would give a total input capacity of 60 if paralleled. If an artificial line with $Z_0 = 150$ ohms is created, the input will be 150 ohms of pure resistance. Such a line must be terminated in Z_0 if it is to be flat. There are several approaches to getting values for L, but since L/C must equal $(Z_0)^2$ we can merely solve

[illegible]

for this relationship, or refer to the chart-figure (6) and pick the values off directly. The intersection of 6 μf with 150 ohms is also the intersection of .15 μh .

The use of line matching sections is not limited to the purely resistive cases. For narrow frequency bands or single frequencies, it is permissible to use the reactive element to be matched as part of the (15) matching section. Thus, if a capacitive reactance is present use it in the L section instead of an external C, or reduce the external C accordingly. If the reactance is inductive, subtract it from the value of X_L to be used in the match. Using this method it is possible to match antennas with widely varying complex impedances to a line or tube.

Division of Complex Numbers

Appendix B

Let $Z_1 =$ Complex no. 1

$Z_2 =$ Complex no. 2

by similar Triangles

$OZ_1'Z_2'$ similar to OZ_2Z_1 ,

but $OZ_1' = 1$

$$\frac{OZ_1'}{OZ_1} = \frac{OZ_2'}{OZ_2}$$

$$\frac{OZ_1'}{OZ_2'} = \frac{OZ_1}{OZ_2}$$

$$\frac{OZ_2'}{1} = \frac{OZ_2}{OZ_1}$$

Normalized

$$|Z_2'| = \frac{Z_2}{Z_1}$$

$$\bar{Z}_2' = Z_2' e^{j(\alpha_1 + \alpha_2)}$$

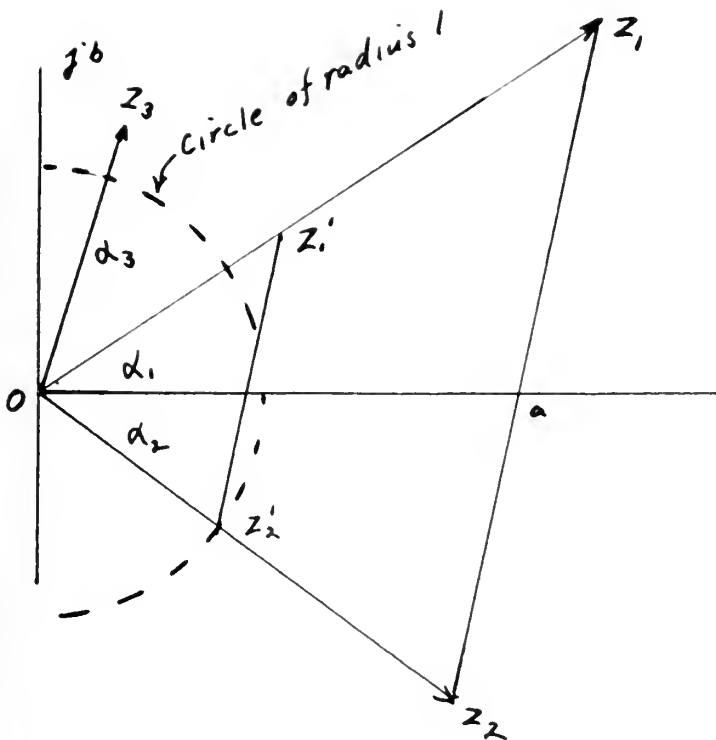


Figure 37

$$G(s) = \frac{R_2}{\left(\frac{L_1}{C} R_2\right) s^2 + \left(\frac{R_2^2}{C} + \frac{L_1}{C^2}\right) s + \frac{R_2^2}{C^2}} = \frac{E_2(s)}{E_1(s)}$$

Let $E_2(s)$ be 1 (see Fig. 13)

Then

$$E_1(s) R_2 = \left(\frac{L_1}{C} R_2\right) s^2 + \left(\frac{R_2^2}{C} + \frac{L_1}{C^2}\right) s + \frac{R_2^2}{C^2}$$

$$E_1(s) = \left(\frac{L_1}{C}\right) s^2 + \left(\frac{R_2}{C} + \frac{L_1}{C^2 R_2}\right) s + \frac{R_2}{C^2}$$

$$E_1(s) = A s^2 + B s + C$$

$$E_1(s) - B s - C = A s^2$$

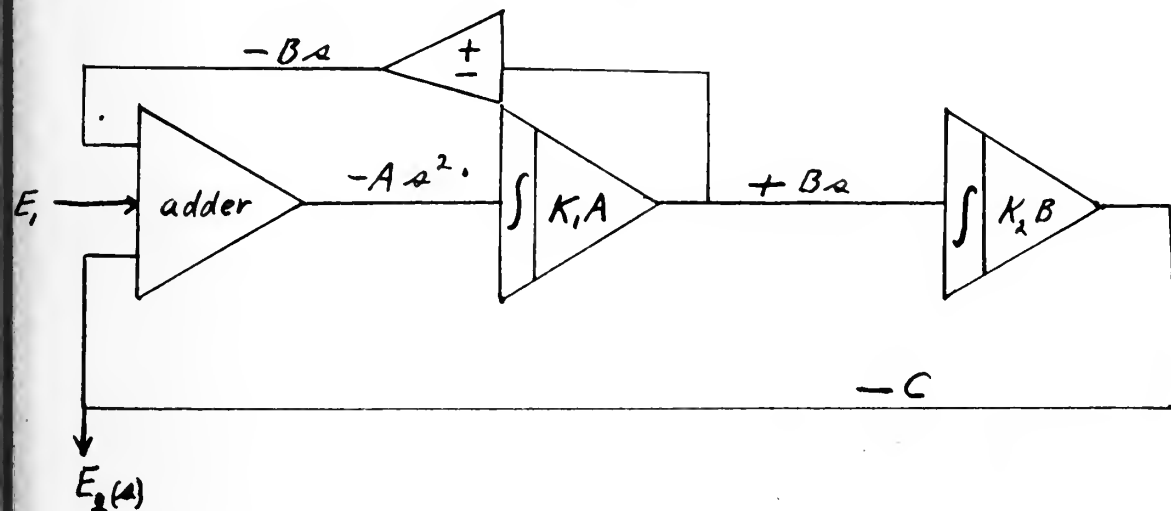


Figure 38 - Analog Computer Data
For the L Match

Let $E_2(s) = 1$ (see Fig. 15)

$$s E_1(s) - B s^2 - C s - 1 = A s^3$$

$$E_1(s) - B s - C - \frac{1}{s} = A s^2$$

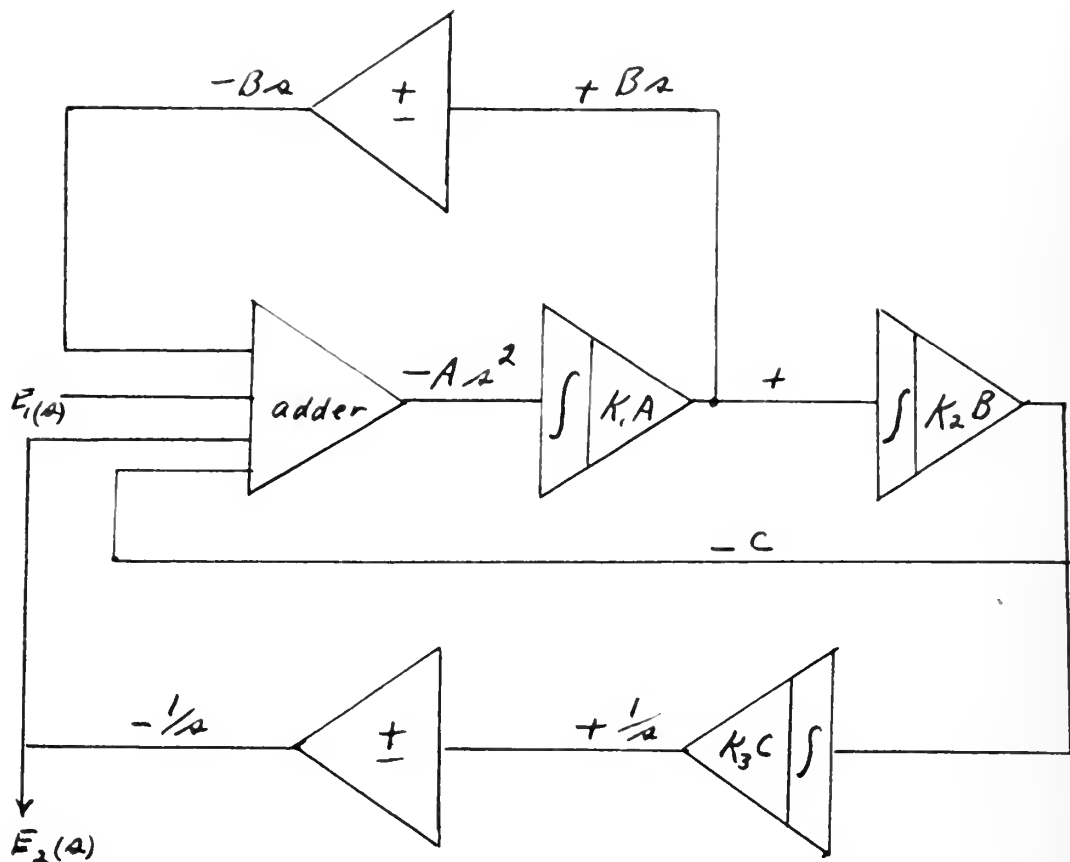


Figure 39 Analog Computer Data
For T Match

$$E_1(s) - Bs^2 - Cs - 1 = As^3$$

(See Fig 14)

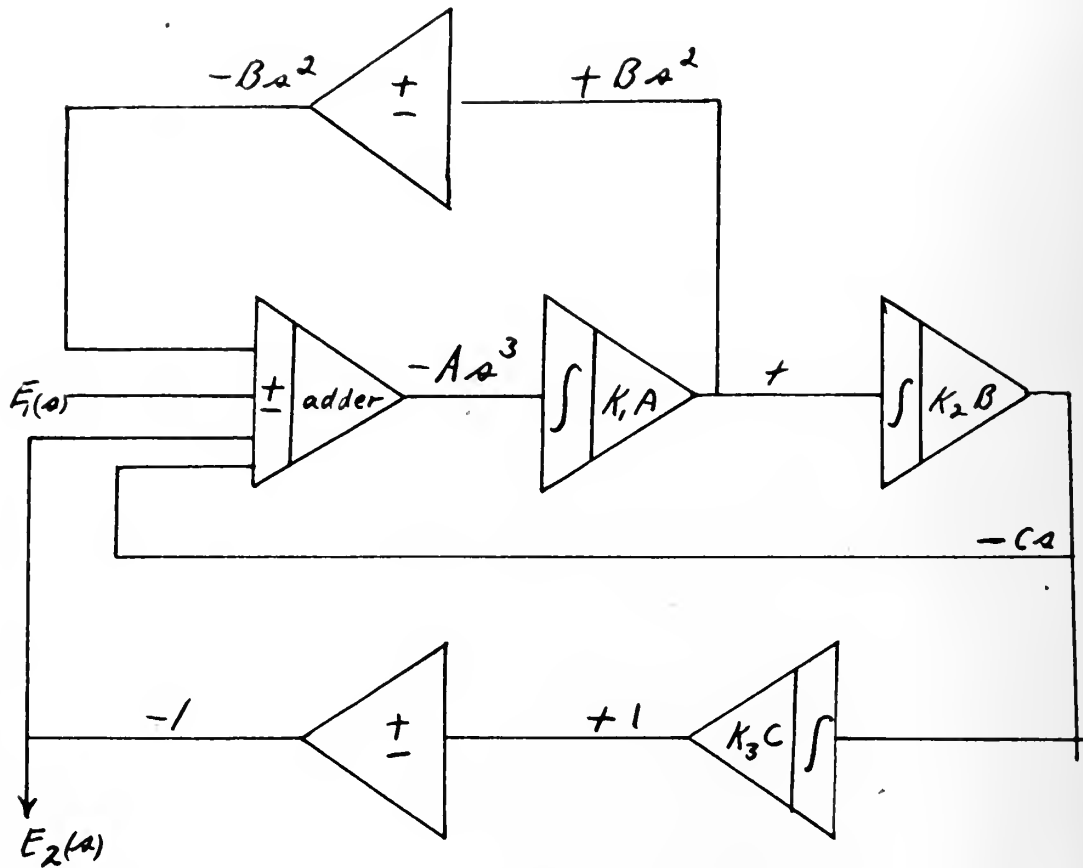


Figure 40 Analog Computer Data
For TT Match.

APPENDIX D

MASTER CAMERA - RECEIVER SYSTEM FOR INDUSTRIAL TELEVISION

The Radio Corporation of America required a master system for use in a customer's plant which would permit an Industrial Television camera to be plugged in at any point, and any number of receivers to be plugged in at other points without interference. It was stipulated that no more than twelve (12) cameras would be operating simultaneously, and each of these could be attached to a small R.F. unit with approximately 2 Volts output feeding a 75 ohm coaxial line. The following solution was proposed by the author, who was then attached to The Radio Corporation of America.

Resistor type drops are used, there being no reason to conserve signal strength. A total attenuation of 50 db is obtained, plus line loss, which still leaves a signal of several millivolts at each receiver.

Loss of signal between points

$$A = 2RL$$

$$\text{Voltage Reflected Ratio } |K| = \frac{\rho - 1}{\rho + 1} = \frac{Z_n - Z_0}{Z_n + Z_0}$$

$$\rho = \frac{Z_0}{Z_n}$$

A slight discontinuity is to be created by
Tapping a coax feed line with a high resistance
takeoff to a receiver - 40db of ghost
rejection is required.

$$\text{if } 2RL = 0$$

$$-40 \text{ db} = \frac{1}{100} \quad \rho \approx .98$$

$$|K| = \frac{1}{100} \approx \frac{.98 - 1}{.98 + 1} = \frac{.02}{1.98} \approx -40 \text{ db}$$

$$\text{for } \rho = .98 \quad Z_n \text{ must be } 50 \times Z_0$$

$$\text{for } 75 \Omega \text{ coax } Z_n = 3,750 \Omega \text{ min.}$$

$$50 \Omega \text{ coax } Z_n = 2,500 \Omega \text{ min.}$$

If $2RL \neq 0$, Z_n may be decreased,
or as is done in practice - more outlets
can be added.

Figure 41 - Development of maximum
Standing Wave Permissible on a
Master Antenna System.

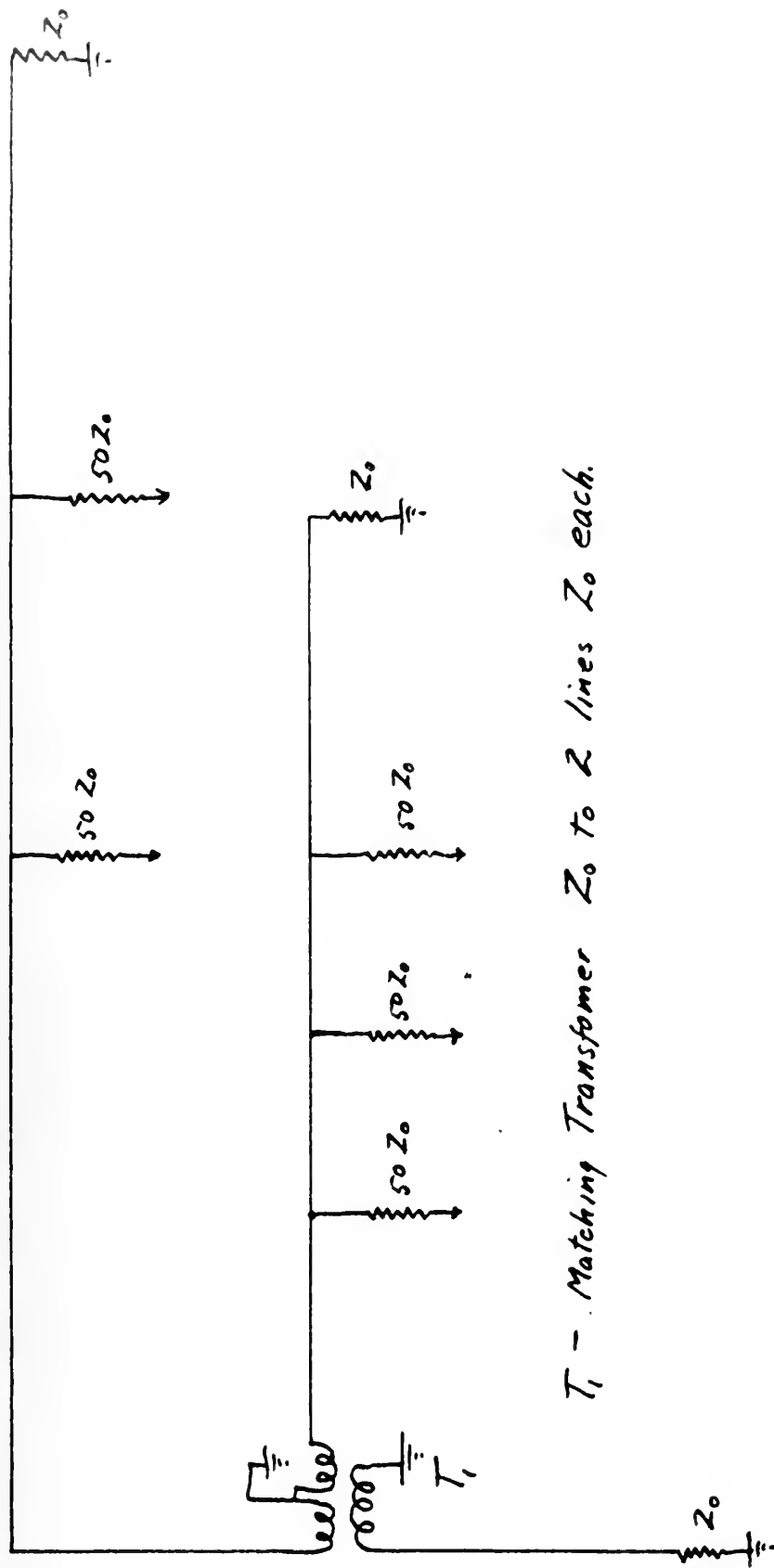
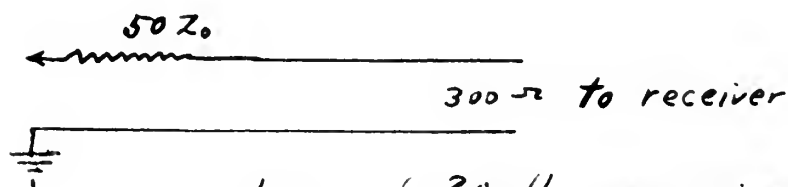
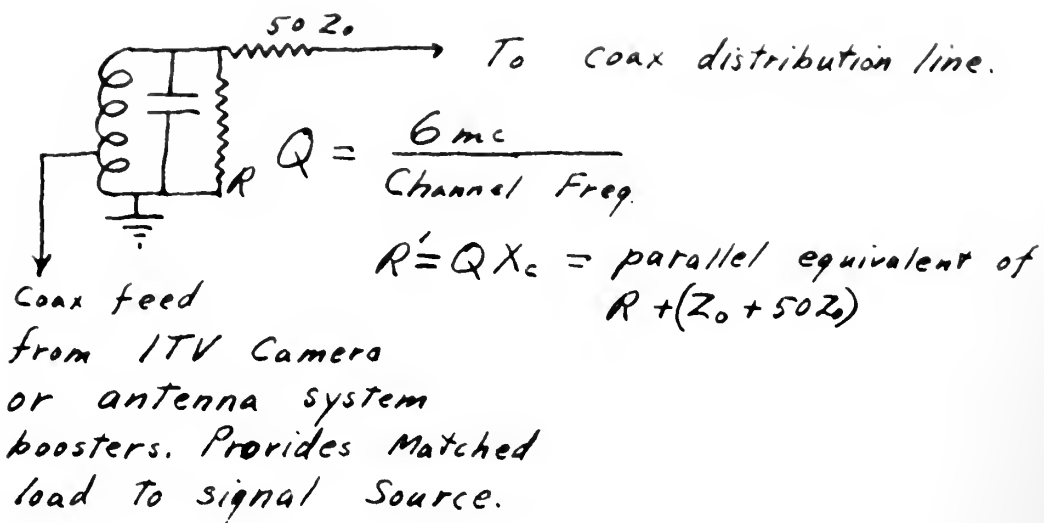


Figure 42 - Master Camera-Receiver Distribution System
for Industrial T.V. and/or master Antenna System
With any number of outlets and Camera
Inputs.



Loss of 20 db approximately
is obtained using this type of
coupling. Signal on line must
be great enough to make up
this loss.



Transformer Coupling for Use with Receiver
Used as line Terminus. This is not
Recommended Because Receivers
Seldom Match at Less than 1.25 VSWR

Figure 43 - Miscellaneous Circuit For
Use with Master Antenna Systems.

RCA - AM-FM-TV Mixer

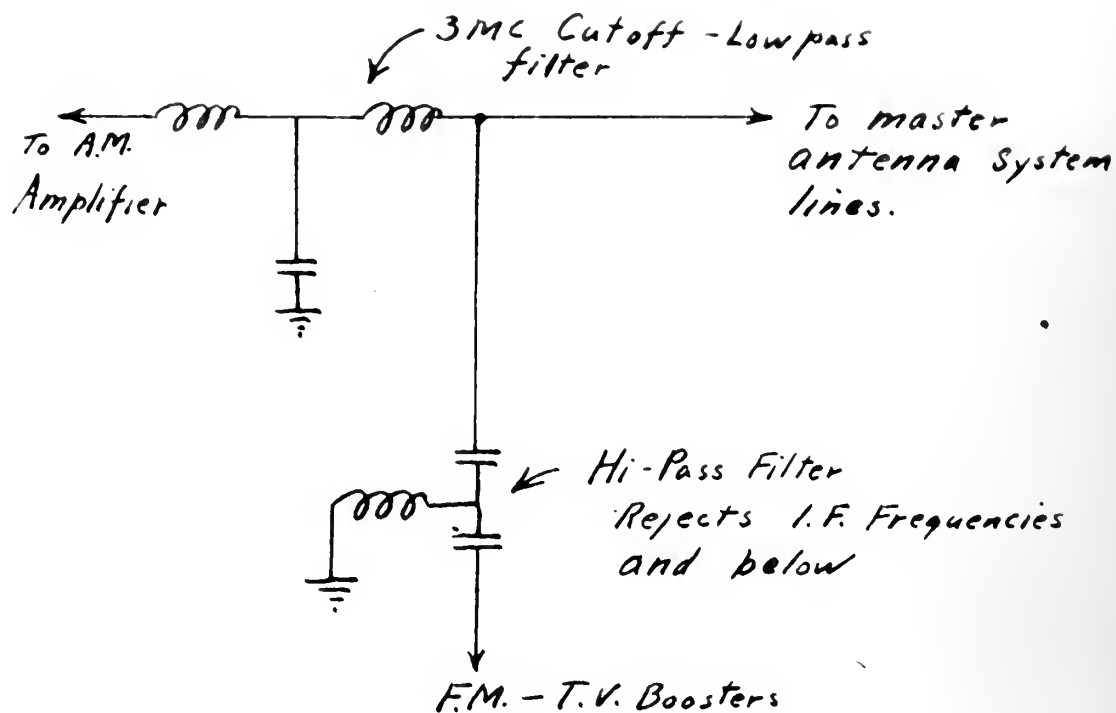
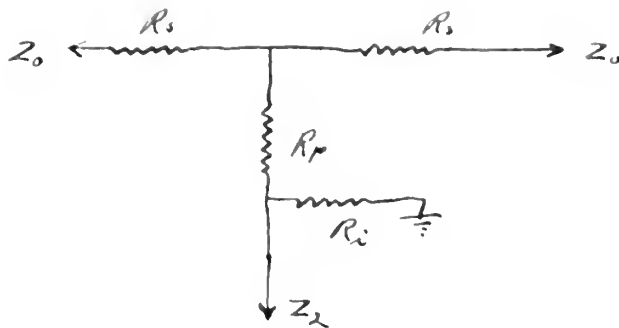


Figure 44 - Miscellaneous Circuit
For Use With Master Antenna Systems.

T Pad to be inserted in Coax Line for perfect match.



$$Z_2 = \frac{R_i (R_p + Z_1 + R_s/2)}{R_i + R_p + Z_2/2 + \frac{R_s}{2}}$$

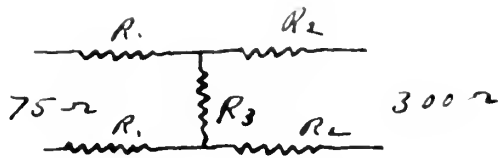
or to a close approximation

$$R_i = \frac{R_p Z_2}{R_p - Z_2}$$

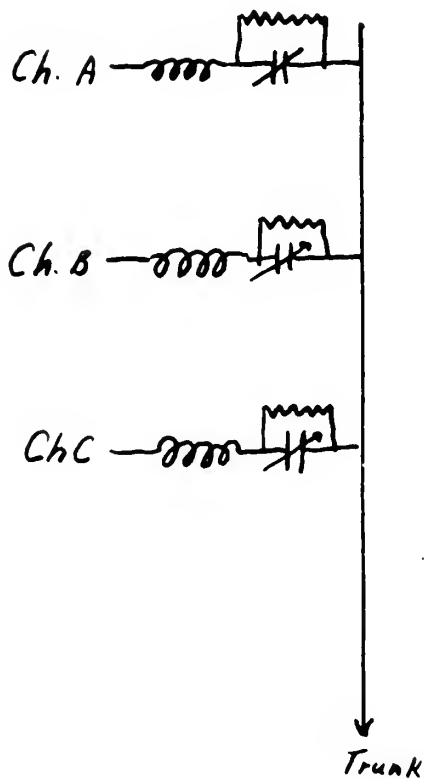
Figure 45



Resistor Pads



Loss db	R_1	R_2	R_3
12 db	0 Ω	130	82
15 db	13	130	56
20 db	24	130	30
25 db	30	130	18



Lynmar Signal
To Trunk Mixer
Feed System.

Figure 46

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• colloidal solution = mixture of two substances in which the particles of one substance are very small and are dispersed in the other substance.

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1. The first step is to identify the problem or question that needs to be answered. This involves understanding the context and the specific requirements of the task.

$\frac{d}{dt} \left(\frac{\partial L}{\partial \dot{x}} \right) = \frac{\partial L}{\partial x}$

1. The first group of people who are not in the labor force are those who are not in the labor force because they are not in the labor force. This group is the largest group of people who are not in the labor force.

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I have been thinking of you a great deal lately and wondering how you are getting on.

It has been a long time since we last spoke and I have been very busy.

I am well and hope these few lines will find you the same.

I am sure you are doing very well and I am glad to hear of your success.

I am sure you are doing very well and I am glad to hear of your success.

I am sure you are doing very well and I am glad to hear of your success.

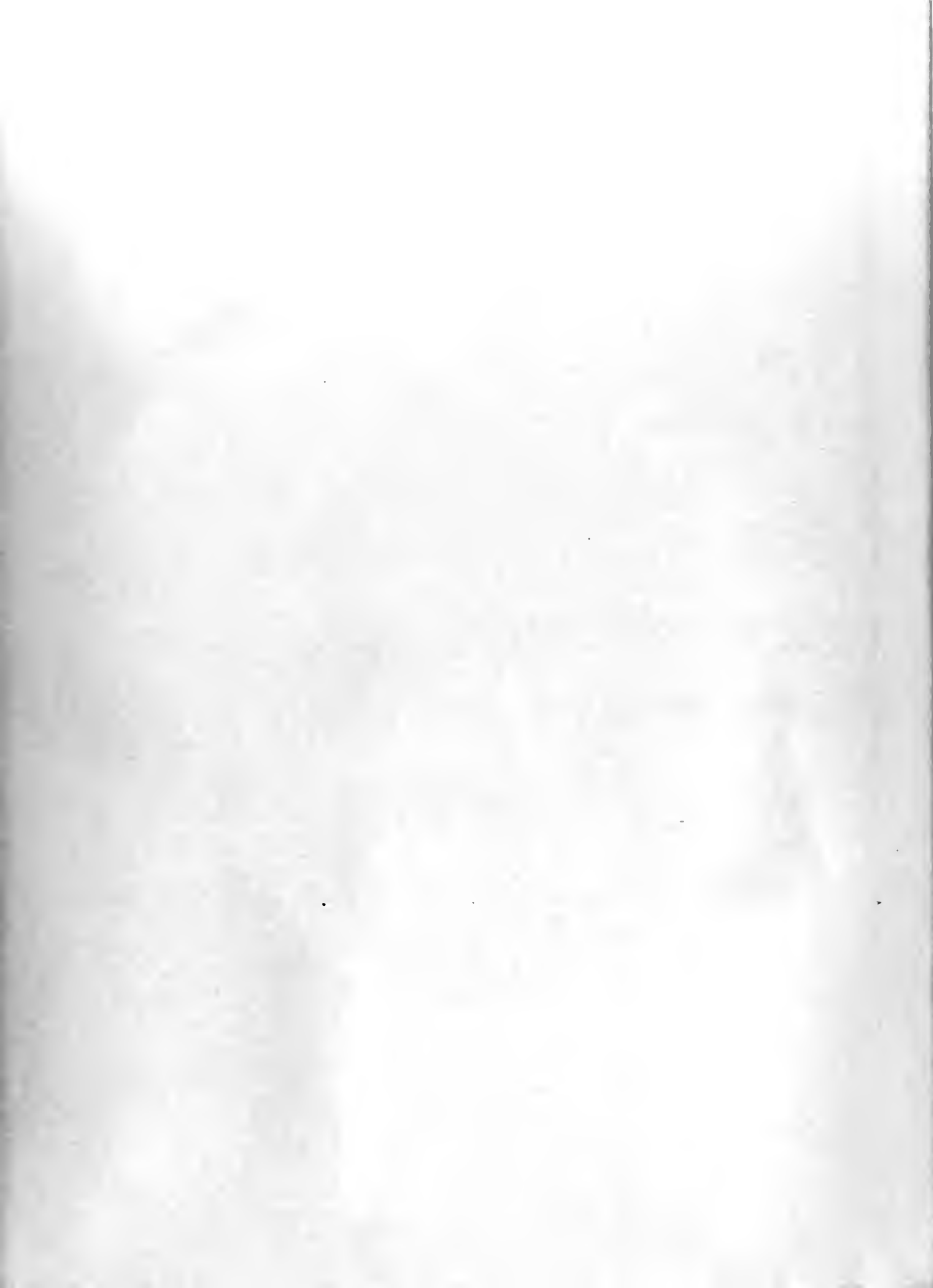
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